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AN ANALYTICAL AND EXPERIMENTAL INVESTIGATION
OF
FM-BY-NOISE JAMMING

THESIS

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OF
FM-BY-NOISE JAMMING

THESIS

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Disclaimer

References to private companies and commercial equipment and software are made in this thesis. All said references are of a strictly academic nature, and no endorsement of any kind is intended by the author or by any person or agency sponsoring or otherwise connected with this research.

Preface

This investigation had two major goals. The first goal was to provide a lucid, complete, and authoritative description of FM-by-noise jamming suitable for open discussion and publication. This first goal has been accomplished. From an analytical perspective, the existing literature on FM-by-noise has been reviewed, consolidated, and highlighted, and, an alternative description of FM-by-noise is proffered. The description of FM-by-noise presented here is significant because it contains reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise at RF and IF. Many symbols and abbreviations are used in this investigation, and a List of Symbols and Abbreviations can be found on page viii.

The second goal of this investigation was to restore the sponsor's ability to measure noise quality and investigate alternative noise quality measures. This goal has been only partially accomplished. Turner *et al.*, in (43), specify the equipment and suggest an algorithm to measure noise quality. Similar equipment has been gathered and the algorithm has been programmed. Additionally, I have proposed two alternative noise quality measures. One measures noise quality at RF, while the other measures noise quality at IF. Further testing of all three noise quality measures just mentioned is necessary.

I am deeply indebted to the members of my thesis committee. Dr. Vittal Pyati and Mr. Eugene Sikora gave me the idea for this thesis. Major Mark Mehalic provided invaluable comments; I owe him many thanks. Dr. Brahmanand Nargarsenker was there when I needed him. Outside my thesis committee, Capt. Jim Hird, my friend and fellow student, listened to my many thesis-related ramblings. Mr. Jim Mudd and Mr. Marvin Potts also provided much support, both moral and technical. Dr. Spyridon Therianos, without whom I would not be an engineer, taught me the beauty and power of mathematics. Numerous engineers from the Hewlett-Packard Company answered my numerous questions about their test equipment. I thank them all. Finally, I thank my beloved wife, Yon Hui, and my children for their years of sacrifice and support.

Charles J. Daly

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List of Symbols and Abbreviations

a .... unnamed ratio defined in (8) similar to the SVR

B .... signal bandwidth

B_{FM} .... bandwidth of FM signal based on Carson's Rule

B_N .... noise bandwidth

B_V .... victim receiver bandwidth

CF .... center frequency

D .... deviation ratio, ratio of Δf_p to bandwidth of modulating signal

Δf_p .... peak frequency deviation

DINA .... direct noise amplification, a jamming technique

D_{RMS} .... RMS deviation ratio, ratio of RMS frequency deviation to modulating signal bandwidth

DVR .... peak frequency deviation-to-victim receiver bandwidth ratio, deviation-to-victim ratio, $\Delta f_p / B_V$

ECCM .... Electronic Counter-countermeasures

ECM .... Electronic Countermeasures

EW .... Electronic Warfare

f_c .... carrier frequency or center frequency

f_d .... frequency-deviation constant in Hz/volt

FM .... frequency-modulation

FM-LFN .... FM-by-low frequency noise, NVR < 1

FM-UBN .... FM-by unity frequency noise, NVR = 1

FM-WBN .... FM-by-wideband noise, NVR > 1

Hz .... Hertz

IF .... intermediate frequency

k_f .... radian frequency deviation constant

LO .... local oscillator

NBFM	narrowband FM
NVR	noise bandwidth-to-victim receiver bandwidth ratio, noise-to-victim ratio, B_N/B_V
pdf	probability density function
p_f	frequency domain penalty
PSD	Power Spectral Density
p_t	time domain penalty
RBW	resolution bandwidth
RF	radio frequency
ρ_{IF}	IF noise quality
RMS	root mean square
ρ_{RF}	RF noise quality
S	sweep rate in Hz/sec as defined in (8)
s	second
SNR	signal-to-noise ratio
SP	spanwidth
SVR	sweep-to-victim receiver bandwidth ratio, sweep-to-victim ratio, SVR = DVR $\cdot$ NVR
SWP	sweep speed
UDRFM	unity deviation ratio FM
V	volt
VBW	video bandwidth
WBFM	wideband FM
ω_c	carrier frequency or center frequency

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Abstract

Among the jamming waveforms used in Electronic Warfare (EW), FM-by-noise is perhaps the least understood, and no exhaustive analytical and experimental investigation of the subject exists. This investigation presents a thorough review, consolidation, and elucidation of the theory on FM-by-noise. To better explain and predict the behavior of FM-by-noise at RF and IF, three useful factors, namely the deviation-to-victim ratio (DVR), the noise-to-victim ratio (NVR), and the sweep-to-victim ratio (SVR), are developed. To complement the theory, results of FM-by-noise jamming experiments conducted using commercial test and measurement equipment are presented. Specifically, the time- and frequency-domain behavior of FM-by-noise jamming at RF and IF, in terms of the DVR, NVR, and SVR, is studied with the baseband noise bandwidth, peak frequency deviation, and victim receiver bandwidth as parameters. An important feature of the experimental portion of this investigation is the inclusion of computer-generated reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise at RF and IF. The behavior of FM-by-noise illustrated in the oscilloscope and spectrum analyzer displays is related back to the description of FM-by-noise jamming developed in the analytical portion of the investigation. Finally, the concept of noise quality, a measure of noise jamming effectiveness, is reexamined. Three noise quality factors are investigated. The work of Turner *et al.* is verified, and two alternative noise quality factors at IF and RF are proposed. *IF noise quality* measures the time-domain normality and frequency-domain uniformity of receiver noise at the output of a victim receiver subjected to FM-by-noise jamming. *RF noise quality* measures the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF. Procedures and programs necessary to duplicate the experiments and noise quality measurements are included.

An Analytical and Experimental Investigation of FM-by-Noise Jamming

I. Preliminaries

Among the jamming waveforms used in Electronic Warfare (EW), FM-by-noise is perhaps the least understood, and no exhaustive analytical and experimental investigation of the subject exists. This investigation presents a thorough review, consolidation, and elucidation of the theory on FM-by-noise. To better explain and predict the behavior of FM-by-noise at RF and IF, three useful factors, namely the deviation-to-victim ratio (DVR), the noise-to-victim ratio (NVR), and the sweep-to-victim ratio (SVR), are developed. To complement the theory, results of FM-by-noise jamming experiments conducted using commercial test and measurement equipment are presented. Specifically, the time- and frequency-domain behavior of FM-by-noise jamming at RF and IF, in terms of the DVR, NVR, and SVR, is studied with the baseband noise bandwidth, peak frequency deviation, and victim receiver bandwidth as parameters. An important feature of the experimental portion of this investigation is the inclusion of computer-generated reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise at RF and IF. The behavior of FM-by-noise illustrated in the oscilloscope and spectrum analyzer displays is related back to the description of FM-by-noise jamming developed in the analytical portion of the investigation.

Finally, the concept of noise quality, a measure of noise jamming effectiveness, is reexamined. Three noise quality factors are investigated. The work of Turner *et al.* is verified, and two alternative noise quality factors at IF and RF are proposed. *IF noise quality* measures the time-domain normality and frequency-domain uniformity of receiver noise at the output of a victim receiver subjected to FM-by-noise jamming. *RF noise quality* measures the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF. Procedures and programs necessary to duplicate the experiments and noise quality measurements are included.

1.1 Background

Electronic Warfare (EW) may be defined as any military action taken to deny enemy forces the effective use of the electromagnetic spectrum and their electronic weapons while simultaneously permitting friendly forces unrestricted use of the electromagnetic spectrum and their electronic weapons. Modern EW is a diverse military science, and one important area of this science is Electronic Countermeasures (ECM). An effective weapon in the arsenal of ECM techniques is active noise jamming. Because of its effectiveness, active noise jamming is the most common form of ECM (33:10).

Active noise jamming seeks to deny enemy forces the effective use of their electronic weapon systems, such as radars, missile guidance systems, and command and control links, by overwhelming these systems with noise. In short, the jamming transmitter *jams* the victim receiver with noise. Noise is the fundamental physical phenomenon limiting the sensitivity of all receivers. Specifically, radar and communications receivers are designed to operate at or above a given signal-to-noise ratio (SNR). Below this threshold SNR, the receiver is unreliable at best or inoperable at worst. The goal of noise jamming then is to intentionally inject as much noise as possible into an enemy force's electronic weapons of war (40:548).

In the literature, noise jamming is classified as either spot jamming, broad-band barrage jamming, or swept-spot jamming. The spot jamming technique employs a jamming transmitter bandwidth that is approximately equal to the bandwidth of the victim receiver. In contrast to spot jamming where jamming power is confined to a relatively narrow bandwidth, broad-band barrage jamming distributes the available jamming power over a wide bandwidth. The two methods represent a trade off between power and bandwidth. Spot jamming delivers high levels of noise power to a relatively narrow band of frequencies, while barrage jamming delivers reduced levels of noise power to a broad band of frequencies. Swept-spot jamming is a combination of spot and barrage jamming. With this hybrid technique, a high level of noise power is swept across a broad band of frequencies (34:14–15).

FM-by-noise jamming, the topic of this thesis, can be used in any of the modes just described. Even though the distinction between spot and barrage jamming is becoming increasingly less important because of advances in the art of ECM, FM-by-noise is often classified as a barrage jamming technique. As is

conventional in the literature, the terms *FM-by-noise jamming* and *FM-by-noise* are used interchangeably throughout this investigation (8:14-9).

Interest in FM-by-noise as a jamming waveform began, like most modern EW topics, with World War II. Papers and reports devoted to FM-by-noise and textbooks containing information or sections on the subject began appearing in the early 1950's. The literature is mentioned in passing here. A detailed literature review, including references, is presented in Chapter II. The early papers and texts on FM-by-noise are sketchy and confusing for variety of reasons, such as unsophisticated measurement equipment, and, perhaps most importantly, the strict security requirements of the times. Since theoretical work on FM-by-noise started in the early 1950's and continued up until the early 1970's, the theory on FM-by-noise is extensive.

In contrast to the extensive theory on FM-by-noise, experimental work on the subject has been limited and sporadic. Two experiments were conducted to determine the effectiveness of FM-by-noise as a jamming waveform. One experiment was conducted in the early 1960's and another was later conducted in 1977 (8; 43). Finally, a computer simulation used to optimize FM-by-noise jamming design, in terms of noise quality, was proposed in 1985 (20).

1.2 Scope

The first of the two goals of this investigation was to explain what FM-by-noise is and how it works. Discussions of other types of jamming and other areas of EW, such as repeater jamming and Electronic Counter-countermeasures (ECCM), did not help to explain what FM-by-noise is or how it works. Hence, discussions on topics such as those just mentioned are not included in this report.

The second goal was to measure the effectiveness of noise jamming. In support of this goal the concept of noise quality as a measure of noise jamming effectiveness is reexamined. The noise-quality work of Turner *et al.* is verified, and two alternative noise quality measures are proposed. While predictions, based on the results presented in this work, about the effectiveness of FM-by-noise against conventional receivers and modulations are justified, predictions about its effectiveness against more sophisticated receivers and exotic modulations should be independently verified.

1.3 Assumptions

This thesis assumes no specific jamming problem or scenario. Originally, a radar jamming problem was assumed; however, equipment limitations forced the use of frequencies and bandwidths more representative of a communications jamming scenario than a radar jamming scenario. Nonetheless, the results obtained are applicable to most FM-by-noise jamming scenarios.

Two other assumptions are noteworthy. In all analytical discussions of FM-by-noise, it is assumed that the center frequency of the jammer and the center frequency of the victim receiver coincided. Furthermore, this first assumption guided the experiments. Specifically, all experiments were conducted with the jammer center frequency being equal to, or at least within 0.5 to 1 kHz of, the center frequency of the victim receiver. Advances in ECM technology are making this assumption increasingly valid and realistic. As will become clear in the sequel, extrapolation to the case where the jammer frequency is appreciably different than the center frequency of the victim receiver is straightforward.

The last assumption concerns the relationship between stationary random processes and histograms. Specifically, the probability density function of the amplitudes of a random process can be estimated from a histogram of amplitude values of a sample function if, and only if, the random process is stationary and the process values are independent for large separations of time (39:192). It was assumed that these conditions, along with the conditions for ergodicity, were satisfied. This assumption was particularly important in measuring noise quality and seemed justified since the behavior of FM-by-noise, at both IF and RF, is essentially time-invariant in the statistical sense (7:342).

1.4 Overview

Despite the extensive documentation on the theory of FM-by-noise, much of it is scattered, disjoint, and abstruse. And, as already alluded to, results from experimental work on FM-by-noise in the open literature are lacking. Chapters II and III address the analytical portion of the investigation. Chapter II presents a thorough review of the open literature and also serves a primer on the theory of FM-by-noise. Chapter III reviews and consolidates the existing theory on FM-by-noise jamming by highlighting the most im-

portant results that have been developed since the 1950's. Chapter III also presents a detailed description of FM-by-noise, paying particular attention to coherence and lucidity. Chapters IV and V constitute the experimental portion of the investigation and serve to complement the analysis of FM-by-noise.

Specifics on experimental limitations and experimental procedure are discussed in Chapter IV. The experiments studied the time-and frequency-domain behavior of FM-by-noise jamming at RF and IF as the baseband noise bandwidth and peak frequency deviation of the FM-by-noise and the bandwidth of the victim receiver were varied. The results obtained from the experiments are presented in Chapter V. A significant feature of this thesis is the inclusion of computer-generated reproductions of the actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise.

Chapter VI reexamines noise quality and proposes two alternative noise quality measures. The work of Turner *et al.* is verified, and two alternative noise quality factors at IF and RF are proposed. *IF noise quality* measures the time-domain normality and frequency-domain uniformity of receiver noise at the output of a victim receiver subjected to FM-by-noise jamming. *RF noise quality* measures the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF. Procedures and programs necessary to duplicate the experiments and noise quality measurements are included. Complete listings of the programs written for this investigation are contained in Appendix A, while Appendix B lists the equipment and software used to conduct the experimental and noise quality portions of this investigation.

II. *Literature Review of FM-by-Noise*

When one researches and reviews EW literature, one is immediately confronted with three problems. First, much of the literature is classified. Second, timely open-literature is usually indirect and abstruse, a direct result of the trade-off between security and academic freedom. Finally, declassified EW literature is often difficult to understand because it was originally written by cleared writers specifically for a cleared audience in a mutually-understood, classified setting. This thesis seeks to provide a lucid, complete, and authoritative description of FM-by-noise jamming suitable for open discussion.

The present chapter is divided into four sections. The first section reviews papers about or directly related to FM-by-noise, while the second section discusses texts about or directly related to the subject. Most of the literature on FM-by-noise makes mention of uniform power spectral density (PSD) barrages. The third section addresses the topic of uniform PSD barrages. Finally, the fourth section summarizes the chapter. Since this chapter provides an in-depth review of the literature on FM-by-noise jamming, it also serves as a primer on FM-by-noise and, thus, provides a point of reference for a more rigorous discussion of FM-by-noise theory which is presented in the next chapter.

2.1 Papers

Beginning in the early 1950's and continuing into the late 1960's, numerous papers and reports directly related to FM-by-noise were published in the open-literature. A representative sampling of the papers and reports written during the period is reviewed (1; 9; 10; 11; 25; 27; 41). The various authors discuss randomly modulated signals from an ostensibly academic point of view and avoid explicit references to EW, *jamming*, and *FM-by-noise*.

Middleton, in (1:690), makes one of the few references to *frequency-modulated by noise* and remarks to his readers that frequency modulation with normal random noise is an approximate model of frequency modulation with speech but acknowledges the existence of more appropriate models. Despite explicit references to EW, jamming and FM-by-noise in these early works, it can be readily assumed that EW,

jamming, and FM-by-noise were the *de facto* topics of the papers and reports published in the 1950's and 1960's, particularly when one considers the affiliations and sponsors of the various authors cited.

Much of the open-literature published during the 1950's was concerned with deriving closed-form or limiting expressions for spectra of signals frequency modulated with random processes (25; 27; 41). Another recurring theme concerned the contribution of low-frequency components of a modulating random process to the overall spectrum of a given FM signal (11; 25: 41). Among these early works, Stewart's report on the power spectra of signals phase- and frequency-modulated by Gaussian noise (41) proved to be the most readable and relevant in terms of the work at hand.

The open-literature published in the 1960's served to complement, refine, and extend the theory developed in the 50's. In 1963, Abramson defined an RMS bandwidth for angle-modulated waves and, subsequently, derived a remarkably simple closed-form expression for the RMS bandwidth of an angle-modulated signal (1). Blachman and McAlpine discussed Woodward's theorem and derived an upper bound on the error associated with this theorem (9). Woodward's Theorem states that the RF spectrum of a wideband FM signal takes on the approximate shape of the probability density of its instantaneous frequency which corresponds to the shape of the univariate probability density of the modulating signal.

Probably because a large collection of classified EW reports was declassified in 1973 (12:xi), few papers directly related to FM-by-noise were published in the 1970's and 1980's. In 1979, Cassara *et al.* proposed a technique for generating a jammer signal with a continuous uniform PSD bandlimited over any desired frequency band (14). Weil, writing in (46), responded to (14) and pointed out that the technique had already been proposed twice. It was proposed once in 1955 by Weil himself and was later independently proposed in 1957. Turner *et al.* briefly discuss FM-by-noise in their article on noise quality (43). Another paper on noise quality, written in 1985 by Knorr and Karantanas, contains a readable description of FM-by-noise (20). Other than the noise quality paper by Knorr and Karantanas, no significant unclassified papers or reports on FM-by-noise published after 1980 could be found.

2.2 Texts

It will be convenient to separate texts containing material on or related to FM-by-noise into two categories: statistical communications texts and EW texts. In the statistical communications category, two

texts stand out: one for what it contains, the other for what it does not. The text by Rowe was one of the more complete and scholarly references on angle-modulation that could be found (32). It consolidates and extends the work done on FM-by-noise in the 50's and 60's. If one has problems obtaining the papers on FM-by-noise published during that period, this text is certainly a suitable substitute. The second text illustrates the difficulty in locating timely open-source information on EW and jamming. Raemer, writing in one of the few statistical communications texts to even have the term *jamming* indexed, notes there is a great body of literature on jamming but quickly adds that "... most of it is classified and cannot be discussed in the open-literature" (31:148).

While the texts on statistical communications theory are helpful, one must turn to EW texts for more practical discussions of FM-by-noise. One of the oldest EW texts, *Threshold Signals* by Lawson and Uhlenbeck, contains a brief, but surprisingly direct, description of FM-by-noise:

> Effects of f-m interference are somewhat different from those of a-m interference. The noise amplitude in the receiver is determined by the excursions of the interference signal across the r-f or i-f acceptance band of the receiver. If one assumes that the total frequency excursion is large compared with the bandwidth of the receiver, then the receiver output will contain a number of pulses whose shape in time is similar to the shape of the i-f or r-f bandwidth in frequency and whose amplitudes are relatively constant. These pulses will be repeated at random times. Because of the relatively constant amplitude of these pulses, the effect of the interference is similar to that of highly limited, or clipped, a-m noise. A "ceiling" effect occurs but for a quite different reason from that in the a-m case.
>
> The effectiveness of interference of this type clearly depends upon obtaining a noise function such that an excursion across the receiver band occurs within a time p approximately equal to the receiver response time, that is, a time equal to the reciprocal of its bandwidth. If such an excursion does not occur, the interference will lose its effectiveness because of the constant-amplitude pulses produced and because of the time spaces between them. Within these spaces the desired signal can be found without any accompanying interference. (22:147)

As mentioned earlier, a large collection of classified EW reports was declassified in 1973. Five years later, these reports were published in complete, unabridged form in the important EW text, *Electronic Countermeasures* (12). Of particular relevance to this investigation are the contributions by Benninghof *et al.* (8) and Morita and Rollins (26) with the former being the most complete and authoritative description of FM-by-noise and its effects that could be found.

Benninghof *et al.* categorize two types of FM-by-noise: *FM-by-wideband noise* (FM-WBN) and *FM-by-low frequency noise* (FM-LFN). They define the two types of FM-by-noise qualitatively, mathematically, and numerically. Qualitatively, they differentiate between FM-WBN and FM-LFN on the basis of modulating noise bandwidth. The authors write,

Frequency modulation by WB noise attempts to produce the same result as does DINA ... It is of interest, however, to investigate the mechanism by which FM by noise techniques can be used to produce jamming that is essentially indistinguishable from DINA [direct noise amplification] at the output of a given radar receiver, and to determine the requirements that must be placed on the FM modulation parameters. Each time the frequency modulated carrier sweeps across the victim's passband, the victim receiver's filter circuits are set to "ringing" by the impulsive character of the input. If the modulation is random, then the receiver input is a random time series of short pulses. If, further, the average frequency of these pulses is much greater than the victim bandwidth, the conditions for the Central Limit Theorem are approximated and the output of the receiver filter (usually i-f) is nearly gaussian in its first order statistics. Thus, one expects the i-f output for FM-by-WB noise to be the same as DINA.

Frequency modulation by LF noise ... restricts the modulating noise bandwidth to be much less than the victim receiver bandwidth. Thus, the ringing caused by one receiver crossing is usually over before another crossing occurs. The i-f output wave is therefore a random time series of distinct pulses whose duration is approximately the reciprocal bandwidth. This jamming, when directed against search radar, exhibits two principal advantages and one principal disadvantage. It has increased effectiveness because the ordinary radar second detector produces more video power for a given i-f output (or receiver input) than with FM-by-WB noise or DINA. Thus, this source is more efficient in producing video jamming than are the others. Also, a given video power is more effective in jamming small targets on a PPI if FM-by-LF noise is used. This may be associated with a confusion effect caused by the resemblance of many of the bright spots to small target echoes. The principal disadvantage of FM-by-LF noise is that it is relatively easy to counter, since the jamming is discontinuous even at the receiver output, and many or most of the target echo pulses are free of jamming if observed in real time. (8:14-10)

FM-WBN applies when the modulating bandwidth is much greater the bandwidth of the victim receiver.

Next, the authors mathematically define FM-LFN and FM-WBN in terms of a linearly swept signal that sweeps across the passband of a receiver. Benninghof *et al.* assume the signal sweeps across the passband with a sweep rate S. This sweep rate will be discussed more fully in the next chapter. The authors provide a detailed mathematical development, which will also be discussed more fully in the next chapter, and write,

... Assuming a barrage width that is large compared to the receiver bandwidth, the sweep speed during any particular transit of the receiver passband by the randomly sweeping jamming will depend, essentially, on the jamming barrage width and the spectral composition of the modulating noise. If the noise spectrum has a sufficiently low upper cut-off frequency (FM-by-LF noise), the receiver output will consist of a series of impulses of random amplitude and random structure. As the modulating noise frequency is increased (FM-by-WB noise), the random impulses will begin to overlap since the pulsewidth depends only on the receiver bandwidth. As a direct consequence of the Central Limit Theorem of probability theory the statistics of the noise so produced at the i-f amplifier will be essentially gaussian. (8:14-21)

For the uninitiated, this passage does little to clarify the complex relationships that exist between modulating noise bandwidth, barrage width or, equivalently, peak frequency deviation, victim receiver bandwidth, sweep rate and receiver output. These complex relationships will be painstakingly described in the next chapter and experimentally illustrated in Chapter V.

Finally, Benninghof *et al.* differentiate between FM-LFN and FM-WBN based on the bandwidth of the modulating noise alone. They assign loose numerical requirements on the bandwidth of the modulating noise for the two types of FM-by-noise. For FM-LFN, the authors require $B_N < 50$ kHz and $B_N > 500$ kHz for FM-WBN. The authors also require that the frequency deviation, peak or RMS is not specified, must be several times greater than the victim receiver bandwidth for both types of jamming (8:14-30). These numerical requirements are mentioned for completeness only. They are not applicable in this investigation. Ultimately, the distinction between FM-WBN and FM-LFN seems to based on the ratio of modulating noise bandwidth to victim receiver bandwidth. Indeed, this interpretation is not only consistent with the terminology but also intuitively appealing. Hence, the terms *FM-WBN* and *FM-LFN* will be borrowed for this investigation and made more precise in the next chapter.

The ratio on which FM-by-noise is categorized can easily be confused with an important ratio from traditional FM theory, namely the deviation ratio. The deviation ratio is defined as the ratio of carrier frequency deviation to modulating signal bandwidth; this ratio will be discussed more fully in the next chapter. Adding to the potential confusion is that fact that this ratio of carrier frequency deviation to modulating signal bandwidth has spawned at least six terms: *slow frequency deviation* and *fast frequency deviation* (22), *high-index FM* and *low-index FM* (9), and, finally, *wideband FM* and *narrowband FM* (47:186). The terms *wideband FM* (WBFM) and *narrowband FM* (NBFM) are the most common today. Note that when discussing carrier frequency deviation, some authors use peak carrier frequency deviation, while others prefer to use RMS carrier frequency deviation.

It must be emphasized that the terms *FM-LFN* and *FM-WBN* and the terms *narrowband FM* and *wideband FM* describe two completely different ratios. Specifically, *FM-LFN* and *FM-WBN* are EW terms that categorize the numerical ratio of modulating noise bandwidth to victim receiver bandwidth. On the other hand, *narrowband FM* and *wideband FM* are terms from basic FM theory that describe the ratio of carrier frequency deviation to modulating signal bandwidth.

Morita and Rollins discuss FM-by-noise in terms of the ratio of modulating noise bandwidth to RMS frequency deviation (26:12-3). This ratio is the inverse of that defined in Stewart's work (41). Oddly enough, Morita and Rollins cited (41) but chose to invert the ratio contained therein. Morita and Rollins

continue and assert that FM-by-noise jamming is more uniform when the ratio of modulating noise bandwidth to RMS frequency deviation is approximately 3.5, but (41) does not support this assertion. Stewart does remark that, for modulating signals that can be modeled by a white Gaussian process, the narrowband FM assumption obtains when the ratio of RMS frequency deviation to modulating bandwidth is less than 0.751 and the wideband FM assumption holds for ratios greater than 0.751 (41:16).

Two additional EW texts contain descriptions of FM-by-noise. The description presented by Lothes *et al.* in (23) is readable and complete. Their description of FM-by-noise begins by considering a noise voltage that is used to frequency modulate a voltage-controlled oscillator (VCO) which has a given VCO constant in Hz/V; the output of the VCO is, in this case, FM-by-noise. They proceed by considering the effect white Gaussian noise has on the output of the VCO under wideband FM conditions. They define the output bandwidth of the VCO as the VCO constant in Hz/V times the RMS value of the noise in volts. They conclude,

> ... we can make this bandwidth [VCO output bandwidth] arbitrarily large by making the noise voltage at the VCO input larger and larger. However, [for a fixed modulating noise bandwidth] the resulting signal is not high quality noise. The VCO may be viewed as a frequency-hopped signal that only rarely "visits" a particular frequency. Although there is power at each frequency in the band, it has an impulsive nature and lacks quality. To get high-quality noise using FM, we want the amount of frequency deviation at the output and the bandwidth of the noise modulation at the input to be about the same ... (23:50–52)

In contrast to the lucid explanation of FM-by-noise presented in (23), the description found in the text by Maksimov *et al.* is most perplexing (24). Their description of FM-by-noise at RF is quite understandable; however, their discussion of its effects at IF is confusing. It will not be repeated here.

Three other EW texts proved helpful. First, Golden describes how a frequency-modulated noise spectrum jams search and scanning radars (18:76–77). Next, Schleher's text (33) provides background material on EW, but its description of FM-by-noise reiterates that presented by Benninghof *et al.* (8). Schleher also discusses a method of generating a uniform PSD jamming barrage. Finally, Schlesinger's text discusses EW from an systems engineering point of view (34).

2.3 Note on Uniform PSD Barrages

Mention must be made of an understandable concern, evident in the literature, with uniform power spectral density (PSD) jamming barrages (14; 8:14-23–14-24; 33:131; 46). It is important to note however

that a barrage with a uniform PSD does not necessarily produce effective jamming. Specifically, Benninghof *et al.* cite work done by Middleton and Weil, who is referred to as Wild by the authors of (8), to modify the shape of an FM-WBN barrage for a more uniform PSD (8:14-23–14-24, 14-61). The process is called *erfing*. Erfing transforms the voltage amplitude distribution of the modulating noise to a uniform distribution. Noise so modified is known as *erfed* or *erfer* noise. The noise is erfed *before* it is used to frequency-modulate the carrier. An FM-by-noise barrage generated with erfed noise has a uniform PSD.

However, Weil, who along with Middleton and others originally developed the erfing process, explains that the statistics of erfed noise are significantly altered by the erfing process and "... although Erfer noise has a uniform PDF, it turns out that it results in a poor distribution of sweep rates and relative number of crossings per second at points away from the center of the voltage waveform (or of the resulting frequency spectrum)" (46:1370). Indeed, Benninghof *et al.* tested barrages generated with erfed noise and found for "... FM sources, the jamming effectiveness is not changed by using 'erfed noise' to 'whiten' the jamming spectrum, as suggested" (8:14-30).

Weil offers alternatives to overcome the problems associated with erfed noise. He writes,

> To overcome [the problems associated with erfed noise] a random triangular wave could be considered. However, this waveform is not truly random; the sweep rate and crossings per second are interdependent, and the crossings tend to come in pairs away from zero voltage.
>
> A random sawtooth waveform solves the problem of pairs but aggravates the other problem [of poor distribution of sweep rates etc.] and is far more difficult to use.
>
> A combination of a high-frequency periodic wave and superimposed (additive) random noise provides a compromise waveform with a relatively uniform PDF and number of crossings per second throughout most of its range, and is probably adequate for most applications. (46:1370)

Finally, Weil describes another alternative which he refers to as *folded-noise*. The details of the folded-noise method can be found in (46).

2.4 Summary

This chapter reviewed the literature on FM-by-noise. The first section reviewed papers and reports about or directly related to FM-by-noise, while the second section discussed texts about or directly related to the same. Finally, the third section discussed methods of generating uniform PSD barrages and the corresponding problems and trade-offs associated with each method.

III. Theory on FM-by-Noise

The descriptions of FM-by-noise jamming referenced in the previous chapter, taken as a unified body of literature, lack consistency in both terminology and development. This chapter develops a description of FM-by-noise based on traditional FM theory. Additionally, a consistent and well-defined terminology is employed. Terms, symbols, and equations are defined as they occur in the development.

3.1 Frequency Modulation (FM) (47:173–188)

3.1.1 General. Consider the familiar general angle-modulated signal given by

$$s_\phi(t) = A_c \cos[\omega_c t + \phi(t)] \tag{3-1}$$

where

A_c = amplitude of the angle - modulated signal in volts

ω_c = center frequency of the angle - modulated signal in radians per second

The instantaneous phase of the general angle-modulated signal is defined as

$$\theta_i(t) = \omega_c t + \phi(t) \tag{3-2}$$

and the instantaneous frequency is defined as

$$\omega_i(t) = \frac{d\theta_i}{dt} = \omega_c + \frac{d\phi}{dt} \tag{3-3}$$

The functions $\phi(t)$ and $d\phi/dt$ are known as the *instantaneous phase deviation*, and *instantaneous frequency deviation* respectively. The instantaneous phase deviation is the quantity of interest in phase-modulation and will not be discussed here. The instantaneous frequency deviation is the quantity of interest in FM, and it will be discussed at length. Note that the value of the instantaneous frequency is additively determined by the value of the carrier frequency and the instantaneous frequency deviation.

Furthermore, the instantaneous frequency deviation of the carrier, about the center frequency ω_c, is proportional to the modulating signal, $m(t)$, and can be written

$$\frac{d\phi}{dt} = k_f m(t) \tag{3-4}$$

where k_f is the *radian frequency deviation constant* in radians per second per unit of $m(t)$, generally volts.

It will be convenient to define a *frequency deviation constant* in Hz per unit of $m(t)$, say f_d, where

$$k_f = 2\pi f_d \tag{3-5}$$

The use of the symbols ω and f for frequency in radians per second and frequency in Hz, respectively, will make the distinction between the units clear throughout the sequel. Frequency in Hz can be assumed when the distinction is not important nor explicitly made. Similarly, the use of the symbols W and B for bandwidth in radians per second and bandwidth in Hz, respectively, will make the distinction between bandwidth units clear; bandwidth in Hz can be assumed when the distinction is not important nor explicitly made. Additionally, all bandwidths, unless explicitly stated or defined, are assumed to be 3 dB bandwidths.

Now, the phase deviation of a frequency-modulated carrier is

$$\phi(t) = 2\pi f_d \int_o^t m(x)\,dx + \phi_o \tag{3-6}$$

and the output of a frequency modulator may be written

$$s_{FM}(t) = A_c \cos\left[\omega_c t + 2\pi f_d \int_o^t m(x)\,dx\right] \tag{3-7}$$

One physical realization of a frequency modulator is the VCO. Unfortunately, there is an upper bound on the amount of frequency deviation a generic VCO can generate. In the absence of modifications or special-purpose equipment, this limit is usually 5-10% of the carrier frequency, f_c (16:31). This limitation was encountered in the experiments and will be commented upon in the next chapter.

Two examples of frequency deviation will prove illustrative. Let k_f equal ω_c radians per second per volt and $m(t)$ be the unit-step function in volts. Using Eqs. 3-3 and 3-4, the value of the instantaneous frequency is simply ω_c for values of time, t, less than zero and $2\omega_c$ for values of time greater than or equal to zero. Next, consider the case when $m(t)$ is a sinusoid. For the purposes of this second example, the exact value of the frequency deviation constant is not important. In this case, the instantaneous frequency will vary periodically about the carrier center frequency in a sinusoidal fashion. This is illustrated in Figure 3-1 where $m(t)$, the slowly varying sinusoid in the figure, is superimposed on the output of a theoretical frequency modulator. Note that the instantaneous frequency has a maximum when $m(t)$ is a maximum and has a minimum when $m(t)$ is a minimum.

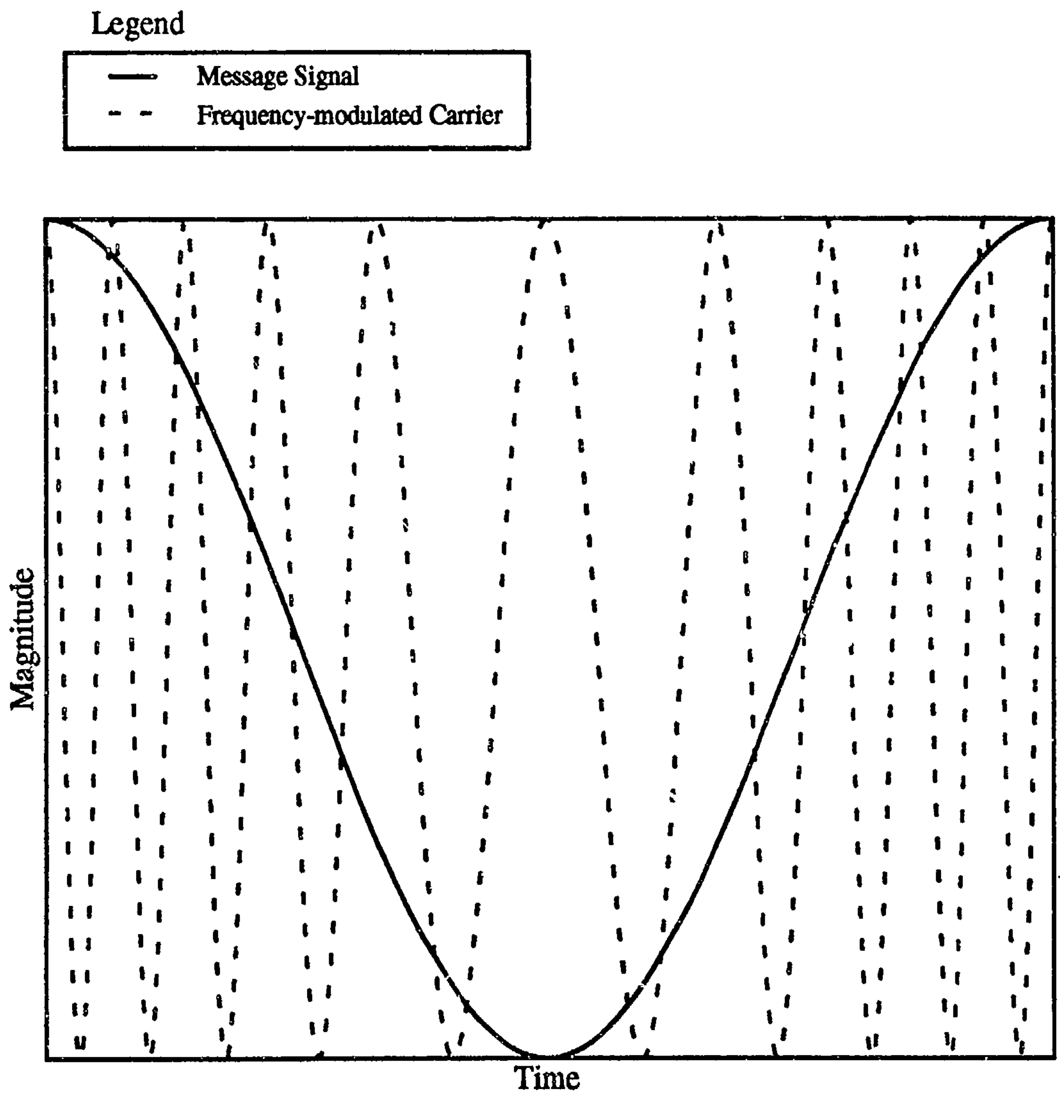


Figure 3-1 Example of Carrier Being Frequency-modulated by a Slowly-varying Sinusoidal Message

3.1.2 FM Bandwidth. Before turning to FM-by-noise, however, FM bandwidth and spectra will be defined. Since the FM process is non-linear, no simple relation exists for relating the spectrum of an FM signal to its *baseband* modulation, except for certain special cases. Hence, most traditional developments of FM bandwidth and spectra start with the special case of sinusoidal modulation and apply the results obtained to approximate more general situations. The development of FM bandwidth based on sinusoidal modulation and Bessel functions is ubiquitous in the literature and will not be repeated here.

As an aside, the term *baseband* implies the notion of data or information residing at its natural location in the frequency spectrum. Modulating an RF carrier with information or data translates the data from its natural or *baseband* location in the frequency spectrum to a new, usually higher, location in the frequency spectrum so that the data can be easily transmitted.

Following Ziemer and Tranter (47), the view is taken that the familiar FM modulation index, β, is defined only for sinusoidal modulation, and, for arbitrary $m(t)$, a commonly used expression for bandwidth results if the *deviation ratio D* is defined as

$$D = \frac{\text{peak frequency deviation}}{\text{bandwidth of } m(t)} \tag{3-8}$$

which is

$$D = \frac{f_d\left[\max\left|m(t)\right|\right]}{B} \tag{3-9}$$

and where it is assumed that B is the 3 dB bandwidth of the baseband signal. The peak frequency deviation will be denoted by the symbol Δf_p. Using Carson's Rule, then, the bandwidth of an FM signal modulated with arbitrary $m(t)$ is

$$B_{FM} = 2(D+1)B \tag{3-10}$$

From Eq. 3-9, it is clear that the deviation ratio can theoretically take on any value from zero to infinity. For $D \ll 1$, B_{FM} is approximately equal to $2B$; this condition is known as *narrowband FM* (NBFM). For $D \gg 1$, B_{FM} is approximately $2DB$ or simply $2 \cdot \Delta f_p$; this is *wideband FM* (WBFM). Note that the bandwidth for WBFM is approximately twice the peak frequency deviation. The FM bandwidth defined in

Eq. 3-10 is not necassarily a null-to-null bandwidth, although in some specific cases, B_{FM} may turn out to be a null-to-null bandwidth. For the purposes of this thesis, the term *unity deviation ratio FM* (UDRFM) refers to any deviation ratio that is equal to one.

It will be convenient to define an *RMS deviation ratio*, D_{RMS}, as

$$D_{RMS} = \frac{\text{rms frequency deviation}}{\text{bandwidth of } m(t)} \tag{3-11}$$

The RMS frequency deviation, left for the moment undefined, will be defined later for the special case when the modulating signal can be modeled by a zero-mean Gaussian process. As a matter of completeness, mention must be made of an ***RMS bandwidth*** which for FM is defined as $1/2\pi$ times the RMS frequency deviation (1:409). Note that the RMS bandwidth as defined in (1) is, in general, not a 3 dB bandwidth.

3.1.3 FM Spectra. As already mentioned, no simple relation fully characterizing the PSD or RF spectrum of an FM signal in terms of its baseband modulation exists, except for certain special cases. The spectrum resulting from frequency-modulating a carrier with a sine wave consists of a series of spectral lines of various amplitudes. The spread of the spectral lines depends on the frequency deviation constant while the spacing between the spectral lines depends on the frequency of the baseband sinusoid. The results are well-documented in the literature and will not be repeated here.

Deriving the PSD or RF spectrum for the case of a carrier frequency-modulated by an arbitrary, random $m(t)$ is no simple task. Woodward's theorem, however, tells us that the RF spectrum of a WBFM signal takes on the approximate shape of the probability density of its instantaneous frequency. Because the instantaneous frequency is proportional to $m(t)$, via Eqs. 3-3 and 3-4, the probability density of the instantaneous frequency has the same shape as the probability density of $m(t)$. Hence, the RF spectrum of a WBFM signal takes on the approximate shape of the univariate probability density function of the modulating signal $m(t)$ (9).

That Woodward's Theorem is true can be shown by considering a carrier that is frequency modulated with a wide-sense stationary random process, $m(t)$ (16:91). It is assumed that the process is zero-mean. The instantaneous frequency deviation at any time t is given by Eq. 3-4. The probability that the in-

stantaneous frequency of the FM carrier lies in the interval $(f, f + df)$ is the same as the probability that $m(t)$ lies in the interval $(f/f_d, f/f_d + df/f_d)$. Using these facts and applying the Mean Value Theorem yields

$$\begin{aligned}
\text{Prob}\left[f < f_i(t) < f + df\right] &= \text{Prob}\left[f - f_c < f_d m(t) < f - f_c + df\right] \\
&= \text{Prob}\left[\frac{f - f_c}{f_d} < m(t) < \frac{f - f_c}{f_d} + \frac{df}{f_d}\right] \\
&= \text{Prob}\left[a < m(t) < b\right] \\
&= \int_b^a p_m(x)\,dx \\
&= p_m(f) \cdot \frac{df}{f_d}
\end{aligned} \tag{3-12}$$

where

$$p_m(x) = \text{probability density function (pdf) of } m(t)$$

and

$$\frac{f - f_c}{f_d} < f < \frac{f - f_c}{f_d} + \frac{df}{f_d}$$

Now, the fraction of the total FM carrier power contained in df is

$$\begin{aligned}
S(f)\,df &= \frac{A_c^2}{2} \cdot \text{Prob}\left[f < f_i(t) < f + df\right] \\
&= \frac{A_c^2}{2} \cdot p_m(f) \frac{df}{f_d}
\end{aligned} \tag{3-13}$$

where $S(f)$ is the PSD or spectrum of the FM signal. Therefore,

$$S(f) = \frac{A_c^2}{2 f_d} \cdot p_M(f) \tag{3-14}$$

Thus, Woodward's Theorem is proved. Note that integrating either side of Eq. 3-14 over the bandwidth of the FM signal yields the total FM carrier power. With basic theory of FM well in hand, attention may now be turned to FM-by-noise jamming.

3.2 A Description of FM-by-Noise Jamming

The terms, symbols, and equations developed in the presentation of traditional FM theory will be helpful in describing FM-by-noise. The description of FM-by-noise developed in this investigation starts by considering the salient features of an FM-by-noise barrage at RF and ends by describing the effects of FM-by-noise at the output of an IF filter. Along the way, various terms, equations, and symbols will be defined and, where possible, compared and contrasted with the existing literature.

3.2.1 FM-by-Noise at RF. FM-by-noise jamming is simply the result of modulating an RF carrier with noise. Usually, the baseband noise, $n(t)$, is white Gaussian noise. Mathematically, we may write a time-domain description of an FM-by-noise barrage, $j_{FM}(t)$, as

$$j_{FM}(t) = A_c \cos\left[\omega_c t + 2\pi f_d \int_o^t n(x)\,dx\right] \tag{3-15}$$

The shape of the resulting RF spectrum or barrage depends on the deviation ratio. Stewart (41:13–16) has shown that the shape of the spectrum is approximately normal, or bell-shaped, for values of RMS deviation ratio much greater than 1. This result is consistent with Woodward's theorem. For values of $D_{RMS} << 1$, however, the barrage is similar to the output of single-tuned resonant circuit driven with white Gaussian noise. The barrage, $S(\Delta\omega)$, for the NBFM case may be written

$$S(\Delta\omega) = \frac{A_c^2/2}{\pi W_N} \cdot \frac{\pi D_{RMS}^2/2W_N^2}{\left(\pi D_{RMS}^2/2W_N^2\right)^2 + \left(\Delta\omega/W_N\right)^2} \tag{3-16}$$

where

$\Delta\omega$ = difference frequency from the unmodulated carrier
W_N = 3 dB bandwidth of the baseband noise

Furthermore, Stewart showed that the 3 dB bandwidth of an NBFM signal is

$$B_{NBFM} = \pi \frac{\Delta f_{RMS}^2}{B_N} \tag{3-17}$$

and the 3 dB bandwidth of a WBFM signal is

$$B_{WBFM} = \Delta f_{RMS} \cdot (8 \ln 2)^{\frac{1}{2}} \tag{3-18}$$

where Δf_{RMS} is the RMS frequency deviation. Again, Stewart asserts that the NBFM condition obtains

when $D_{RMS} < 0.751$, and the WBFM condition applies when $D_{RMS} > 0.751$. These results apply to the special case of a baseband signal that can be modeled by a white Gaussian process. Note also that Eqs. 3-17 and 3-18 do not represent RMS bandwidths.

It is convenient and appropriate, at this point, to discuss the RMS frequency deviation of a carrier frequency-modulated by a signal that can be modeled by a wide-sense stationary, zero-mean Gaussian process. The assumed Gaussian process is not necessarily white. In this case, the probability density of the signal voltages is approximately normal. Hence, it can be assumed that the maximum or peak voltage of the modulating process is $3 \cdot \sigma$ where σ is the RMS voltage of the process. Therefore, considering the numerator of Eq. 3-9, the approximate peak frequency deviation can be written

$$\Delta f_p = f_d(3\sigma) \tag{3-19}$$

Thus, the RMS frequency deviation is approximately

$$\Delta f_{RMS} = f_d \sigma \tag{3-20}$$

As a result of this observation, the RMS frequency deviation of a FM carrier, frequency-modulated with a zero-mean Gaussian process, can be found from the peak frequency deviation by simply dividing the peak frequency deviation by three. This point is made, and will be returned to, because the peak frequency deviation was the parameter that could be directly controlled with the equipment used in the experiments.

3.2.2 FM-by-Noise Ratios and Terms Defined. Before describing the effects of FM-by-noise at IF, a few terms must be defined. First, it will be convenient to define a *noise-to-victim ratio* (NVR)

$$\mathrm{NVR} = \frac{\text{bandwidth of baseband noise}}{\text{bandwidth of victim receiver}} = \frac{B_N}{B_V} \tag{3-21}$$

and, like the deviation ratio in FM theory, consider three cases. The terms *FM-LFN* and *FM-WBN* will be borrowed from Benninghof *et al.* (8) and made more precise. First, FM-LFN implies a NVR < 1. Next, FM-WBN implies a NVR > 1. The numerical requirements mentioned in (8) for FM-LFN and FM-WBN are not applicable in this investigation. Finally, *FM-by-unity bandwidth noise* (FM-UBN) implies a NVR = 1.

As an aside, the abbreviations used to describe the NVR can be combined with the abbreviations used to describe deviation ratio D. For example, WBFM-LFN describes wideband frequency-modulation

by low frequency noise. Additionally, the bandwidth of the victim receiver is assumed to be equal to the bandwidth of the victim receiver's IF filter. Hence, the terms *victim receiver bandwidth* and *IF filter bandwidth* are used synonymously throughout the sequel.

Next, the *deviation-to-victim ratio* (DVR) is defined:

$$\text{DVR} = \frac{\text{peak frequency deviation}}{\text{bandwidth of victim receiver}} = \frac{\Delta f_p}{B_V} \tag{3-22}$$

Unlike the deviation ratio D in FM theory and the NVR just defined, there is no need to further classify the DVR. The reasons for not classifying the DVR will be commented upon later. Finally, the NVR and DVR can be used to define a *sweep-to-victim ratio* (SVR)

$$\text{SVR} = \text{NVR} \cdot \text{DVR} = \frac{B_N \cdot \Delta f_p}{B_V^2} \tag{3-23}$$

The SVR gives an indication of how fast and how frequently an FM-by-noise jammer sweeps through the passband of the its victim receiver.

The SVR just defined is mentioned in the literature in a variety of guises. Hence, comparing and contrasting the SVR with the existing literature requires considerable discussion. First, Benninghof *et al.* describe FM-by-noise in terms of a signal that is linearly swept through a receiver passband (8:14-20). They assume a receiver filter response that is approximately Gaussian and given by

$$G(\omega) = A_1 \exp\left[\frac{-(\omega - \omega_o)}{2b^2}\right] \tag{3-24}$$

where the 3 dB bandwidth of the receiver is approximately equal to $1.67 \cdot b$. Next, they define a linearly swept signal

$$v_i(t) = A_2 \cos\left(\frac{St^2}{2}\right) \tag{3-25}$$

where S is the sweep rate.

Benninghof *et al.* continue and define an unnamed ratio

$$a = \frac{S}{b^2} \tag{3-26}$$

The authors use a in Eq. 3-26 and develop equations for the output of a filter characterized by Eq. 3-24 when the input is given by Eq. 3-25. They consider two cases: slow sweeps, $a << 1$ and fast sweeps, $a >> 1$. After they consider the two cases, they explain FM-LFN and FM-WBN with the second passage quoted on page 2-4 in the previous chapter.

The sweep-to-victim ratio or SVR defined in Eq. 3-23 can be compared and contrasted with the variable a defined by Eq. 3-26. Comparatively, the two ratios are similar because the numerator of the SVR and the numerator of a can both be interpreted as sweep rates in Hz per second. Despite this similarity, the two ratios are different because a is exact and the SVR is a statistical approximation.

The ratio a is exact because the sweep rate S in Eqs. 3-25 and 3-26 is exact. On the other hand, however, the SVR is an approximation because the numerator of the SVR, $\Delta f_p \cdot B_V$, approximates, in a statistical sense, the average sweep rate of the FM-by-noise barrage. That is, the sweep rate associated with each sweep of the FM-by-noise across the passband of the victim receiver is random specifically because the sweep rate is determined by the structure of each zero-crossing of the baseband noise which is, of course, random. Hence, the sweep rate of the FM-by-noise barrage can be approximated in a statistical sense only. This discussion will be numerically illustrated in Chapter V. For the purposes of this investigation, differences between the transfer function of the actual victim receiver and the transfer function defined by Eq. 3-24 are ignored.

Quantities similar to the SVR are mentioned in two other references. Blachman asserts

> ... the duration of the transient response of the filter must be small compared to the ratio of the filter bandwidth [B_V] to the rate of change of frequency [numerator of SVR, $\Delta f_p \cdot B_N$]. Since the duration of the transient response is of the order of magnitude of the reciprocal of the filter bandwidth, this means that the filter bandwidth must be large compared to the geometric mean of modulation bandwidth [B_N] and frequency excursion [Δf_p]. Of course, the filter bandwidth must be small compared to the frequency excursion ... (10:56)

The correspondence between Blachman's definitions and the terms defined in this chapter have been editorially noted. The SVR is similiarly related to the excursion time, p, mentioned by Lawson and Uhlenbeck (22:147) and discussed in Chapter II.

3.2.3 FM-by-Noise at IF. The effects of FM-by-noise at IF can now be described fairly easily by considering the deviation ratio or D, the deviation-to-victim ratio or DVR, and the noise-to-victim ratio or

NVR. Before proceeding, it must be emphasized that the effectiveness of FM-by-noise jamming cannot be described by a single quantity or ratio. Any meaningful discussion of FM-by-noise must consider, at least, the three ratios just mentioned. The SVR is useful when one wishes to relate the slope of the zero-crossings of the baseband noise to the output of the victim receiver.

FM-by-noise takes a baseband noise signal and translates it to RF while preserving or, more commonly, spreading the baseband noise bandwidth. Note that the baseband noise signal could be transmitted directly as in DINA, but DINA results in amplitude clipping because of power amplifier limitations. In order to preserve or spread the baseband noise bandwidth, D must be at least one. That is, the literature seems to imply that FM-by-noise barrages based on NBFM are not effective barrages. However, as Lothes *et al.* have noted, increasing D for a fixed B_N does not produce a more effective barrage; in fact, they require Δf_{RMS} be approximately one (23:52).

Increasing D, or equivalently Δf_p, while keeping the jamming power and baseband noise bandwidth constant decreases the effectiveness of the resulting barrage in two ways. First, the power-bandwidth trade-off inherent in any real jamming scenario limits the effectiveness of the barrage because increasing the peak frequency deviation spreads the available power out over a wider range of frequencies but proportionally decreases the power at each frequency. Not only does increasing the peak frequency deviation reduce the overall power spectral density, it also forces the frequency modulator to deviate over a wider range of frequencies. Thus, the resulting noise barrage behaves like a frequency-hopped signal that "... only rarely 'visits' a particular frequency" (23:52).

In general, noise barrages are at as least as wide as their intended victims, usually in terms of 3 dB bandwidth. Likewise, in order for FM-by-noise jamming to have a seriously debilitating effect on its victim receiver, the FM-by-noise barrage must be at least as wide as the victim receiver. This means that the DVR must be at least equal to or, preferably, greater than one. This is the reason the DVR is not further classified in the manner D and the NVR are. A large DVR FM-by-noise barrage is not, however, necessarily better than one with a smaller DVR. Increasing the DVR, for a fixed B_V, implies increasing D and contending with the attendant trade-offs.

An optimum value for the DVR, in terms of the victim receiver bandwidth, can be found by considering Eq. 3-18 through Eq. 3-20. Solving Eq. 3-18 for Δf_{RMS} with $B_{WBFM} = B_V$ yields

$$\Delta f_{RMS} = 0.425\ B_V \tag{3-27}$$

Now, using Eqs. 3-19 and 3-20

$$\Delta f_p = 1.27\ B_V \tag{3-28}$$

Therefore, from Eq. 3-22 the optimum DVR, in terms of is B_V, is 1.27.

The next ratio to consider in an FM-by-noise jamming scenario is the NVR and its three classifications: FM-LFN, FM-UBN, and FM-WBN. It is assumed that both the deviation ratio, D, and deviation-to-victim ratio, DVR, are greater than one. Also, it is assumed that the jammer center frequency is the same as the center frequency of the victim receiver. Extrapolation to the case when the jammer is off center frequency is straightforward.

FM-LFN and FM-WBN produce different results at the output of the IF filter of the victim receiver. First, consider FM-LFN. In this case, B_N is less than B_V. Hence, the jammer carrier is only occasionally swept near or through the IF filter. The exact output of the filter depends upon the sweep of the carrier and, of course, the bandwidth and shape of the victim receiver. For complete sweeps across the full bandwidth of the receiver, the output is a sine wave of increasing or decreasing frequency, depending on the direction of the sweep, which is amplitude-modulated by the shape of the filter; the output has a duty cycle corresponding to the length of time it took the carrier to make the full sweep across the full passband of the victim receiver. For sweeps that enter the receiver passband but veer out before completing a full run across the receiver bandwidth, the output is a sine wave, appropriately weighted by the filter transfer function, whose duty cycle corresponds to the amount of time the carrier was actually in the passband of the receiver. The frequency content of the sine wave depends on the exact nature of the carrier's excursion into the passband. This description of FM-LFN is illustrated in Chapter V.

Some authors emphasize the discrete nature of these damped sinusoids and refer to them as pulses (8:14-20: 22:147), while others prefer to highlight the frequency content of the damped sinusoids and refer to them as *rings* or *ringing* (20:274). Both terms will be used throughout this investigation. Because of its

pulse-like nature, FM-LFN produces a confusion effect by causing numerous false targets to appear on a radar scope.

In contrast to FM-LFN, which is crude deception jamming technique, FM-WBN is a noise jamming technique which produces a jamming effect similar to direct noise amplification (DINA). Because FM-WBN relies on higher frequency noise than FM-LFN, the number of full sweeps across the victim receiver passband, which is related to the number of zero-crossings of the baseband noise process, is guaranteed to increase. This statement assumes a fixed receiver bandwidth, B_V. Usually, the receiver does not have time to recover from one sweep before another full sweep occurs. Thus, the receiver responses overlap, and "... [t]he Central Limit Theorem guarantees that the output waveform will become gaussian as the number of overlapping responses becomes large" (20:274), and the receiver is saturated with gaussian noise. Benninghof *et al.* also invoke the Central Limit Theorem to justify this result and specify a loose numerical inequality such that FM-by-noise approximates DINA. They write,

> ... [i]n order [for FM-by-noise jamming] to meet the requirements for producing jamming that is equivalent to DINA, the following inequality must be satisfied $f_R < f_N < f_J$ where f_R $[B_V]$ is the receiver bandwidth, f_N $[B_N]$ is the average noise bandwidth, and f_J is the jamming barrage width. Typical numerical values for the qualities (*sic*) are $f_R = 1$ mc, $f_N = 5$ mc, $f_J = 200$ mc (8:14-22).

Neither reference, however, presents a rigorous proof of their arguments. Indeed, necessary and sufficient conditions, in terms of the NVR, DVR, and SVR, for FM-by-noise to approximate DINA are not well-developed in the literature, nor will they be developed in this investigation. As will be shown later, FM-UBN seems to behave much like FM-WBN. Both FM-WBN and FM-UBN are illustrated in Chapter V.

3.3 Summary

This chapter developed a description of FM-by-noise based on traditional FM theory. First, basic FM theory was discussed. The relationship between baseband voltage, frequency deviation constant f_d and instantaneous frequency was emphasized. Next, the deviation ratio D and peak frequency deviation Δf_p were defined. This led to a definition of FM bandwidth known as Carson's Rule. The deviation ratio was also used to define three types of FM: NBFM, UDRFM, and WBFM. Next, FM spectra were discussed and

Woodward's Theorem, which states that the RF spectrum of a WBFM signal has the approximate shape of the univariate probability density function of its modulating signal, was proved.

Attention was then turned to FM-by-noise. First, FM-by-noise at RF was considered. Following this, numerous ratios and terms were defined before describing the effects of FM-by-noise at IF. These ratios included the noise-to-victim ratio or NVR, the deviation-to-victim ratio or DVR, and, the sweep-to-victim ratio or SVR. The NVR is used to characterize FM-by-noise jamming as FM-LFN, FM-UBN, or FM-WBN. FM-LFN produces a confusion effect, while FM-WBN produces an effect similar to DINA. The effect of FM-UBN as observed in this investigation was similar to that of FM-WBN. The DVR indicates how wide a given FM-by-noise barrage is in comparison to the bandwidth of its intended victim. Finally, the SVR indicates how frequently a given FM-by-noise jammer will sweep through the passband of its victim receiver and how long these sweeps will last. These three ratios taken together can be used to explain and predict the behavior of a given FM-by-noise barrage, and its effectiveness, at IF.

IV. Experiments

Any one of the verbal descriptions of FM-by-noise presented thus far would be more understandable if it were illustrated with meaningful graphics. Unfortunately, few of the original references contain explicit figures to illustrate and supplement their verbal descriptions. This thesis seeks to remedy the situation by providing a lucid and complete description of FM-by-noise, complete with an extensive set of illustrative graphics. The goal of the experiments, then, was to observe and record the salient features of FM-by-noise at both RF and IF, and subsequently, present the results in some graphically meaningful form. Despite severe equipment limitations and constraints, this goal has been accomplished. In fact, the graphs presented in the next chapter are a significant feature of this investigation. This chapter discusses the test equipment and procedures used to obtain the data that is presented in the graphs. Additionally, equipment limitations and engineering compromises are discussed. A complete list of the test equipment used can be found in Appendix B.

4.1 Test Equipment

In order to keep the experiments practical and manageable, commercial test and measurement equipment of a generic variety was used to simulate, observe, and record various FM-by-noise jamming scenarios. The basic set-up consisted of a simulated jammer and victim receiver. A block diagram of the equipment set-up is shown in Figure 4-1. A white Gaussian noise generator connected to the FM input of a signal generator served as a jammer, while a simple bandpass filter simulated the IF filter of the victim receiver. The FM-by-noise barrage at the RF output of the signal generator was mixed down, by a second signal generator which served as a local oscillator (LO), to the center frequency of the bandpass filter and, thus, *jammed* the simulated IF filter. The output of the IF filter was connected to test and measurement equipment.

The test and measurement equipment consisted of the following: a programmable digitizing oscilloscope, a programmable spectrum analyzer, and a computer. The programmable digitizing oscilloscope and programmable spectrum analyzer were used to observe and record the time- and frequency-domain

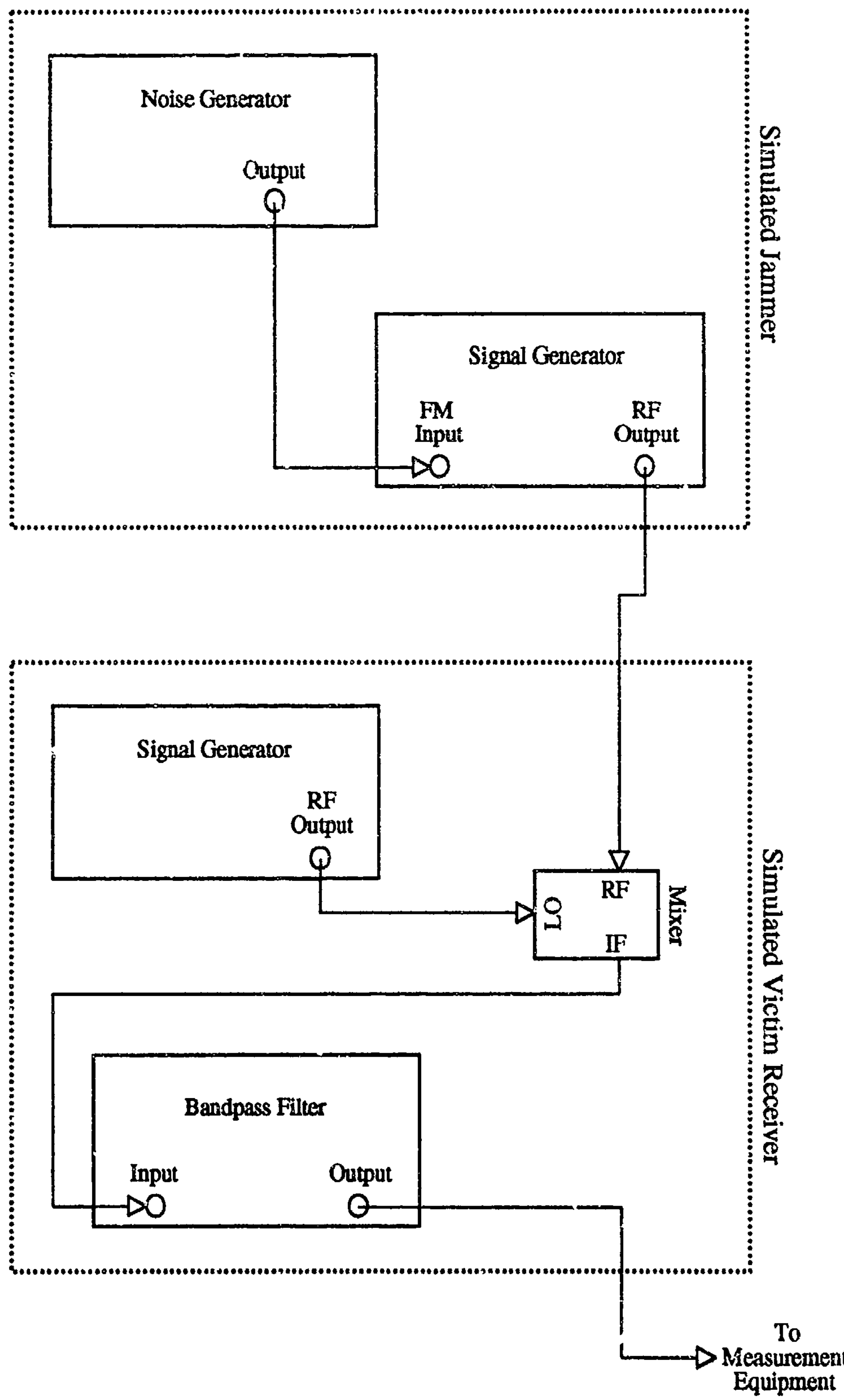


Figure 4-1 Block Diagram of Equipment Set-up

behavior of FM-by-noise at RF and IF. The trace data from the oscilloscope and spectrum analyzer were downloaded to the computer as data files and subsequently plotted using a commercial software package. Hence, the time- and frequency-domain plots presented in the next chapter are computer-generated reproductions of the actual oscilloscope and spectrum analyzer displays observed in the test set-up. A complete listing of the programs used to digitize and download the traces is contained in Appendix A.

Figure 4-2 is a photograph of the equipment set-up. The spectrum analyzer is shown in the bottom left-hand corner of the photograph, and on the shelf directly above the spectrum analyzer, is the noise generator. The bandpass filter is on top of the noise generator. To the right of the bandpass filter is the signal generator that was used as the LO for the simulated victim receiver. Under this first signal generator is an identical signal generator that was used, in conjunction with the noise generator, to simulate a FM-by-noise jammer. The digitizing oscilloscope is below the two signal generators. The mixer is the small three-pronged device resting on the oscilloscope in the photograph. Finally, the computer that controlled the spectrum analyzer and digitizing oscilloscope is shown on the right of the photograph.

While the use of standard test and measurement equipment kept the experiments practical and manageable, it also presented unique engineering challenges and trade-offs. Equipment limitations forced numerous iterations in the development of a workable experimental procedure. Each time a new procedure was developed and tested, previously unknown equipment limitations would halt experimentation and force a reevaluation of the experimental procedure. The equipment limitations, as described in the next section, will convey a sense of the engineering challenges and trade-offs confronted while conducting this investigation.

4.2 Equipment Limitations

4.2.1 Noise Generator. Generating the large noise bandwidths required for noise jamming with commercial equipment was difficult. A wideband noise generator was devised by terminating the input of an amplifier with a resistor and then selecting the highest amplification level available on the amplifier. While this configuration did produce wideband noise, the output power level was too low to drive the FM input of a signal generator. Ultimately, a noise generator with a maximum bandwidth of 50 kHz was used.

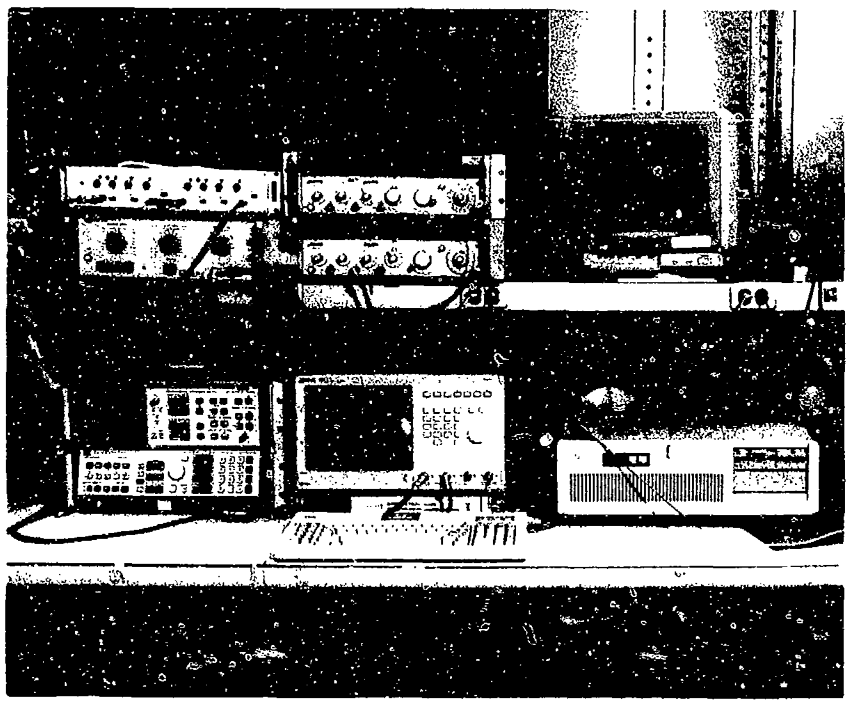


Figure 4-2 Photograph of Equipment Set-up

4.2.2 Signal Generator. Like the noise generator, the signal generator used in this investigation also imposed certain experimental limitations. Three different commercial signal generators were evaluated before one was chosen. None of the signal generators tested was capable of accepting wideband modulating signals while simultaneously tolerating large peak frequency deviations. The signal generator chosen had a 3 dB FM input bandwidth of 250 kHz. That is, the signal generator would have effectively lowpass filtered, to 250 kHz, any wideband (B_N > 250 kHz) signal applied to its FM input. Therefore, even if a wideband noise generator were available, noise bandwidth settings above 250 kHz could not have been selected. The relatively narrow FM input bandwidth of the signal generator was not an issue, however, because the noise generator used in the experiments was limited to 50 kHz.

The signal generator was also limited in terms of absolute maximum peak frequency deviation. Consistent with the physical limitations mentioned in the previous chapter, the maximum peak frequency deviation was limited to 1% of the lowest frequency in a given tuning range. For example, the maximum peak frequency deviation available in the tuning range from 256 to 512 MHz was 2.56 MHz. The highest tuning range extended from 512 to 1024 MHz. Hence, the absolute maximum peak frequency deviation available was limited to 5.12 MHz. Using Eq. 3-10 and assuming WBFM, then, the maximum barrage bandwidth was limited to approximately 10.24 MHz. In short, a commercial signal generator capable of accepting wideband modulation at its FM input while simultaneously tolerating large peak frequency deviations could not be obtained for this investigation.

4.2.3 Digitizing Oscilloscope. An HP 54111D digitizing oscilloscope was used in the experiments. Discussing the limitations of the digitizing oscilloscope now rather than those of the bandpass filters will make the remainder of the chapter more coherent and understandable. The experimental limitations imposed by the digital oscilloscope were formidable especially when one considers the typical FM-by-noise jamming scenario and the goal of the experiments.

The real-time operating mode of the digitizing oscilloscope was limited in terms of sweep speed, sampling rate, memory depth, vertical resolution, and bandwidth. Referring to Table B-2, it can be seen that the sweep speed or, equivalently, the sampling rate determines the memory depth, while the vertical

resolution, in bits, determines the bandwidth. Note that the vertical resolution is either 6, 7, or 8 bits. An additional vertical resolution setting was available. Specifically, the vertical resolution could be set to OFF for quick data-acquisition times.

The memory depth is best explained in terms of a single sweep. The oscilloscope makes a single sweep and takes 8192 samples of the input signal. The sampling rate is determined by the sweep speed, but the number of samples recorded in memory by the oscilloscope is always 8192 samples. Note that the sampling rate times the memory depth is approximately 8192 samples for each sampling rate and memory depth shown in Table B-2. The time interval spanned or represented by these 8192 samples is the memory depth.

Only 501 samples of the total 8192 samples could be displayed on the oscilloscope trace at a given time. Furthermore, data could be acquired from one of two memory partitions in oscilloscope memory: trace memory or channel memory. Trace memory contained the 501 samples displayed on the oscilloscope display, while channel memory contained the full 8192 samples in the full memory depth. The oscilloscope trace displays in Chapter V were plotted with 501 data samples per channel. Data for noise quality measurements discussed in Chapter VI was acquired from channel memory in blocks of 8192 samples.

As evidenced by Table B-2, higher sampling rates could resolve higher frequency signals, but higher sampling rates also resulted in shorter oscilloscope displays and lower memory depths. Increasing the vertical resolution proportionally decreased the maximum resolvable frequency. Finally, the oscilloscope was a lowpass sampling unit; hence, signals could be not be bandpass-sampled without significant pre- or post-processing.

4.2.4 Bandpass Filters. The experimental limitations confronted due to the bandpass filter used in the experiments and the digitizing oscilloscope just discussed were intimately related, a direct consequence of the Sampling Theorem. Two bandpass filter banks were tested. The first unit had eight selectable bandwidths ranging from 0.1 MHz to 23.2 MHz. The 0.1 MHz bandpass filter was centered at 20 MHz while the higher bandwidth filters were centered at 60 MHz. Although these center frequencies and bandwidths are representative of a typical radar jamming scenario, they proved to be too high and fast for

the purposes of this investigation. Even though the oscilloscope could accurately sample, reconstruct, and display signals at these frequencies, the sampling rates required to do so made the sweep times or display lengths too short.

That is, the sweep times resulting from these higher sampling rates were not long enough to simultaneously record both the behavior of the baseband noise used to generate the FM-by-noise barrage and the effect of the barrage at the output of the IF filter of the simulated victim receiver. It was crucial to observe and record the baseband noise because the behavior of the baseband noise can be related to the behavior of FM-by-noise at RF and IF.

Bandpass sampling with no pre- or post-processing was tried but produced unacceptable results. Furthermore, FM-UBN and FM-WBN could not be investigated with this first bandpass filter unit because the noise generator was limited to 50 kHz. Recall that FM-UBN and FM-WBN are based on the NVR, and the noise generator could not produce a signal with a bandwidth equal to or greater than any of filter bandwidths.

The second filter unit tested was a dual low-pass, high-pass filter; the two could be cascaded to form a bandpass filter. This dual filter bank provided a versatile means of forming bandpass filters with tunable center frequency and adjustable bandwidth. The center frequency was tunable from zero to approximately 115 kHz, and the bandwidth was adjustable, subject of course to typical filter constraints such as those cited by Gagliardi (16:155). This unit was chosen for use in the experiments. Two bandpass filters were subjected to FM-by-noise jamming in the experiments. The first filter had a 3 dB bandwidth of 25 kHz and center frequency of 102.5 kHz, and the second had a 3 dB bandwidth of 50 kHz and center frequency of 90 kHz. These values, which are approximate, were chosen so that FM-LFN, FM-UBN, and FM-WBN scenarios could be investigated.

4.3 Engineering Compromises

Considering the typical FM-by-noise jamming scenario and the experimental limitations imposed by the test equipment, it would seem that the experiments were doomed to failure. Happily, such was not the case. In the end, judicious engineering compromises enabled the collection of meaningful data. The ac-

tual bandwidths and center frequencies used are more representative of a communications jamming scenario than a radar jamming scenario. Even though the jamming scenarios used were unrealistic in terms of radar, the behavior of FM-by-noise at RF and IF was accurately simulated, observed, and recorded.

Ultimately, three variables could be fairly well controlled in the laboratory. First, the baseband noise bandwidth, B_N, could be changed with a knob on the noise generator. Three baseband noise bandwidths were used: 5 kHz, 15 kHz, and 50 kHz. Next, the peak frequency deviation, Δf_p, could be varied with a vernier on the signal generator. Nominally, ±1 volt produced the selected peak deviation in the positive and negative directions, respectively. For example, if the carrier were tuned to 100 MHz and the peak deviation were 1 MHz, the FM RF output would alternate between 99 MHz and 101 MHz given that $m(t)$ were a ±1 volt square wave. Finally, the bandwidth of the victim receiver IF filter, B_V, was chosen so that FM-LFN, FN-UBN and FM-WBN scenarios could be investigated.

Since the jamming was mixed down to 90 kHz or 102.5 kHz, large peak frequency deviations could not be used due to problems of frequency fold-over caused by the mixing process. This is best explained by considering the negative frequencies associated with a double-sided PSD. As the frequency-modulated carrier was deviated, by the baseband noise, toward DC, the carrier's image in the negative frequency-domain was also deviated toward DC. If the carrier was deviated too far below DC, its negative counterpart would fold-over into the positive-frequency domain and appear in the filter passband. Consequently, the peak frequency deviation was limited to avoid problems of excessive frequency fold-over.

However, limiting Δf_p too much would result in NBFM. Hence, Δf_p was set at 150 kHz to avoid problems with frequency fold-over while simultaneously maintaining WBFM conditions. Since Δf_p was 150 kHz and the center frequencies of the simulated victim receivers were 90 kHz and 102.5 kHz, respectively, frequency fold-over could not be completely eliminated but only made negligible. This was done by simultaneously reducing the RMS voltage of the noise and re-adjusting the peak frequency deviation on the signal generator. Additionally, the triggering level of Channel One was adjusted as necessary so that Channel One triggered on a baseband noise voltage of approximately 0.3 volts; thus, the corresponding Channel Two trace, which measured the filter output, did not exhibit the effects of frequency fold-over.

4.4 Experimental Procedure

The experimental limitations imposed by the equipment and the engineering compromises described in the previous section dictated an experimental procedure. This section describes how the simulated jammer and simulated victim receiver were set-up and how the measurements were made. Recall Appendix B contains a complete list of the equipment used in the experiments. Although the procedures are specific to the equipment listed in Appendix B, they can be used as a point of reference for a set-up based on different equipment. Familiarity with programmable spectrum analyzers and programmable digitizing oscilloscopes is assumed.

4.4.1 Baseband Noise Measurements. Connect the HP 82324A Measurement Coprocessor, installed in the computer, to the spectrum analyzer and digitizing oscilloscope with HP-IB cables. Daisy-chain the analyzer and oscilloscope. On the HP 3722A Noise Generator select INFINITE SEQUENCE LENGTH and depress RUN. Toggle the switch under the VARIABLE 600 Ω OUTPUT to GAUSSIAN. Select the RMS AMPLITUDE and GAUSSIAN NOISE BANDWIDTH desired. Next, connect the VARIABLE 600 Ω OUTPUT to the spectrum analyzer and Channel One of the digitizing oscilloscope. Finally, run the TIMEDMN.BAS and FREQDMN.BAS programs listed Appendix A to download the time-and frequency-domain representations of the baseband noise. Time- and frequency-domain measurements of 5 kHz and 50 kHz bandlimited, white Gaussian noise were taken for this investigation.

4.4.2 Measurement of FM-by-Noise at RF. Connect the HP 82324A Measurement Coprocessor, installed in the computer, to the spectrum analyzer with an HP-IB cable. There is no need to disconnect the spectrum analyzer from the oscilloscope if the two are already daisy-chained to the computer with HP-IB cables. Set-up the simulated jammer. On the HP 3722A Noise Generator, select the desired baseband noise bandwidth; see Section 4.4.1. Connect the VARIABLE 600 Ω OUTPUT of the noise generator to the FM INPUT of an HP 8640B Signal Generator. The FM INPUT of the HP 8640B has a 3 dB bandwidth of 250 kHz. Therefore the HP 8640B will effectively lowpass filter to 250 kHz any signal, greater in bandwidth than 250 kHz, applied at its FM INPUT.

Connect the RF OUTPUT of the signal generator to the spectrum analyzer. On the signal generator, select the desired frequency range with the RANGE knob. Adjust the COUNTER MODE to INT and

select the desired number of digits to be displayed. Tune to the desired RF using the FREQUENCY TUNE and FINE TUNE verniers. Adjust the OUTPUT LEVEL to 0 dBm; turn the inner vernier on the OUTPUT LEVEL selector fully clockwise. Flip the RF OUTPUT switch on the signal generator to ON. Observe the unmodulated RF carrier on the spectrum analyzer. Phase-lock the carrier to the tuned frequency by depressing LOCK. Adjust the spectrum analyzer settings and OUTPUT LEVEL as required.

Generate an FM-by-noise barrage by modulating the RF carrier with baseband noise from the HP 3722A Noise Generator. Of the four-positions—INT, AC, DC, CAL—on the switch labeled FM, select AC. Depress the FM/kHz button on the SCALE panel. Adjust the PEAK DEVIATION to the desired peak frequency deviation Δf_p. The meter on the SCALE panel indicates the amount of peak frequency deviation. Observe the FM-by-noise barrage on the spectrum analyzer. This is FM-by-noise at RF. Adjust the spectrum analyzer settings as required. Run the FREQDMN.BAS program listed in Appendix A program to download the analyzer trace.

Five FM-by-noise barrages were generated for this investigation. To investigate the impulsive nature of FM-by-noise *in the frequency-domain*, three barrages were generated at 250 MHZ using baseband noise bandwidths of 500 Hz, 5 kHz, and 50 kHz, respectively; all three barrages were generated with a peak frequency deviation of 150 kHz. The impulsive nature of FM-by-noise in the frequency-domain is referred to as *frequency-domain impulsivity*. More will be said about the impulsive nature of FM-by-noise in the next chapter. Finally, to demonstrate the power-bandwidth trade-off inherent in noise jamming, two barrages were generated at 900 MHz using peak frequency deviations of 150 kHz and 4 MHz, respectively; both barrages were generated using baseband noise bandwidths of 50 kHz.

4.4.3 Measurement of FM-by-Noise at IF. Connect the HP 82324A, installed in the computer, to the spectrum analyzer and digitizing oscilloscope with HP-IB cables. Daisy-chain the analyzer and oscilloscope. Set-up the simulated jammer. On the HP 3722A Noise Generator, select the desired baseband noise bandwidth; see Section 4.4.1. Connect the VARIABLE 600 Ω OUTPUT of the noise generator to the FM INPUT of an HP 8640B Signal Generator and Channel One of the oscilloscope.

Set-up the simulated victim receiver. Connect the RF OUTPUT of the first HP 8640B Signal Generator to the connector labeled RF on the ANZAC mixer. Connect the RF OUTPUT of the second HP 8640B Signal Generator to the mixer connector labeled LO. Connect the mixer output labeled IF to the IN 1 input of the Wavetek Rockland 852 Dual Hi/Lo Filter Bank. On the filter bank, connect OUT 1 output to IN 2 input. On filter 1 select HIGH PASS, 0 dB GAIN, and FLAT GAIN. On filter 2, select LOW PASS, 0 dB GAIN, and FLAT GAIN. Connect OUT 2 to the spectrum analyzer and Channel Two of the oscilloscope.

The cascaded filters form a bandpass filter. Set the cut-off frequencies for filter 1 and filter 2 and determine the center frequency of the bandpass filter. Set the jammer carrier frequency to coincide with the center frequency of the simulated victim receiver. This can be done by the following steps. Set the center frequency of the spectrum analyzer to the center frequency of the bandpass filter. While observing the spectrum analyzer display, mix the *unmodulated* jammer carrier down to center frequency of the bandpass filter by adjusting the either the jammer carrier frequency or the LO frequency. Once the jammer carrier is mixed down to the center frequency of the bandpass filter, begin FM-by-noise jamming by selecting the desired peak frequency deviation; see Section 4.4.2. Adjust the spectrum analyzer and oscilloscope settings. Set the oscilloscope for a single sweep by depressing the RUN/STOP button and adjust the trigger level if necessary. Run the program TFDATA.BAS, listed in Appendix A to download time-domain representations of the baseband noise and the filter output and a frequency-domain representation of the filter output. This is the set-up that was used to observe and record the behavior of FM-by-noise at IF *and* to measure noise quality.

The FM-by-noise jamming scenarios investigated are shown in Table 4-1. Note the jamming scenarios are described using the terminology developed in the previous chapter. Since the instantaneous frequency of the barrages was determined by the instantaneous voltage of the baseband noise, $n(t)$, via Eqs. 3-3 through 3-5, the deviation of the baseband noise voltage about zero volts can be interpreted as a direct representation of the deviation of the carrier about its center frequency and the center frequency of the filter. Recall the jammer center frequency coincided with the center frequency of the victim receiver. Therefore, not only was the time-domain output of the IF filters observed and recorded, but the baseband noise

voltage was also observed, recorded and directly related to the output of the IF filters. Thus, a clear picture of FM-by-noise jamming was taken and shall be presented in the next chapter.

Table 4-1 FM-by-Noise Jamming Scenarios Used in Experiments

$B_V = 25$ kHz		$\Delta f_p = 150$ kHz		
B_N	D	NVR	DVR	SVR
0.5 kHz	300	0.02	6	0.12
5 kHz	30	0.20	6	1.20
50 kHz	3	2.00	6	12.0
$B_V = 50$ kHz		$\Delta f_p = 150$ kHz		
B_N	D	NVR	DVR	SVR
0.5 kHz	300	0.01	3	0.03
5 kHz	30	0.10	3	0.30
50 kHz	3	1.00	3	3.00

4.5 Summary

The goal of the experiments in the this portion of this investigation was to observe and record the salient features of FM-by-noise at both RF and IF, and subsequently, present the results in some graphically meaningful form. This chapter discussed the test equipment and procedures used to obtain the data that is presented in the graphs in the next chapter. Additionally, equipment limitations and engineering compromises were discussed. Noise quality measurement were also made, but these measurements are discussed separately in Chapter VI.

V. Results

This chapter contains the results obtained from the experiments described in Chapter IV. Computer generated reproductions of actual oscilloscope and spectrum analyzer displays observed in the laboratory are presented in the figures, and the discussions in the chapter are based on these figures. First, baseband noise will be briefly discussed. Next, the behavior of FM-by-noise at RF is examined. Finally, the effects of FM-by-noise at IF are investigated and explained. The discussions in this chapter rely heavily on the terminology developed in Chapter III. The reader may consult that chapter or the List of Symbols and Abbreviations, on page viii, as necessary.

Several abbreviations are used in the figures. Although the abbreviations used in the figures are contained in the List of Symbols and Abbreviations, they are annotated here for the reader's convenience: CF = *C*enter *F*requency, SP = frequency *SP*an, VBW = *V*ideo *B*and*W*idth, RBW = *R*esolution *B*and*W*idth, and SWP = time to *SW*ee*P* one trace.

5.1 Baseband Noise

Typical time- and frequency-domain representations of the bandlimited, white Gaussian noise generated by the HP 3722A noise generator and used in the experiments are shown in Figures 5-1 and 5-2. Two examples of the baseband noise were deemed sufficient. Note that the 5 kHz noise has, as expected, a smaller bandwidth and less zero-crossings than the 50 kHz noise. Figures 5-1 and 5-2 show no other extraordinary features than those just mentioned. The first two figures are presented for completeness and comparative purposes.

5.2 FM-by-Noise at RF

Figures 5-3 through 5-6 illustrate the behavior of FM-by-noise at RF. Here RF refers to the output of the simulated jammer; see the block diagram in Figure 4-1. All barrages were centered at 250 MHz with the exception of the barrage shown in Figure 5-6 which was centered at 900 MHz.

Consider Figures 5-3 through 5-5 first. The figures all have a similar format. Each figure shows two different displays, acquired from the same spectrum analyzer, of the same FM-by-noise barrage. Dif

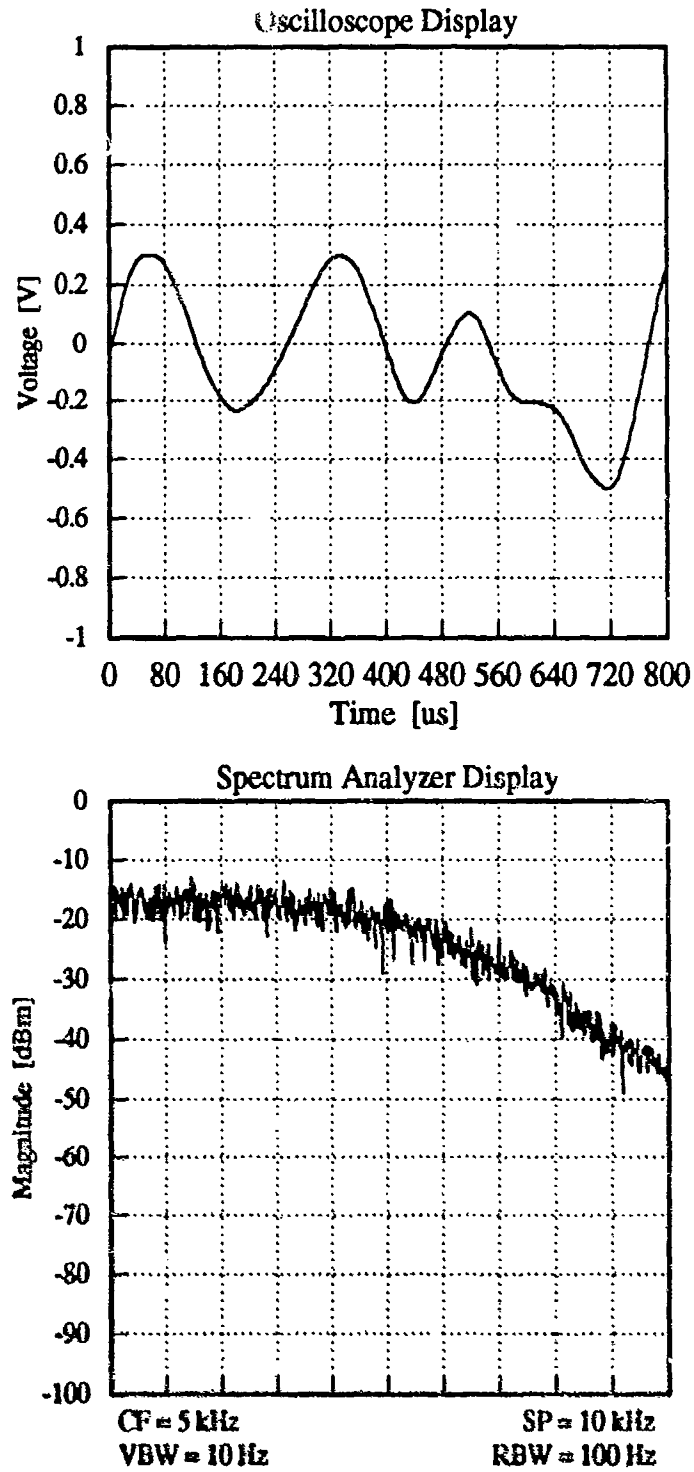


Figure 5-1 Typical Time and Frequency Domain Representations of 5 kHz Bandlimited White Gaussian Noise

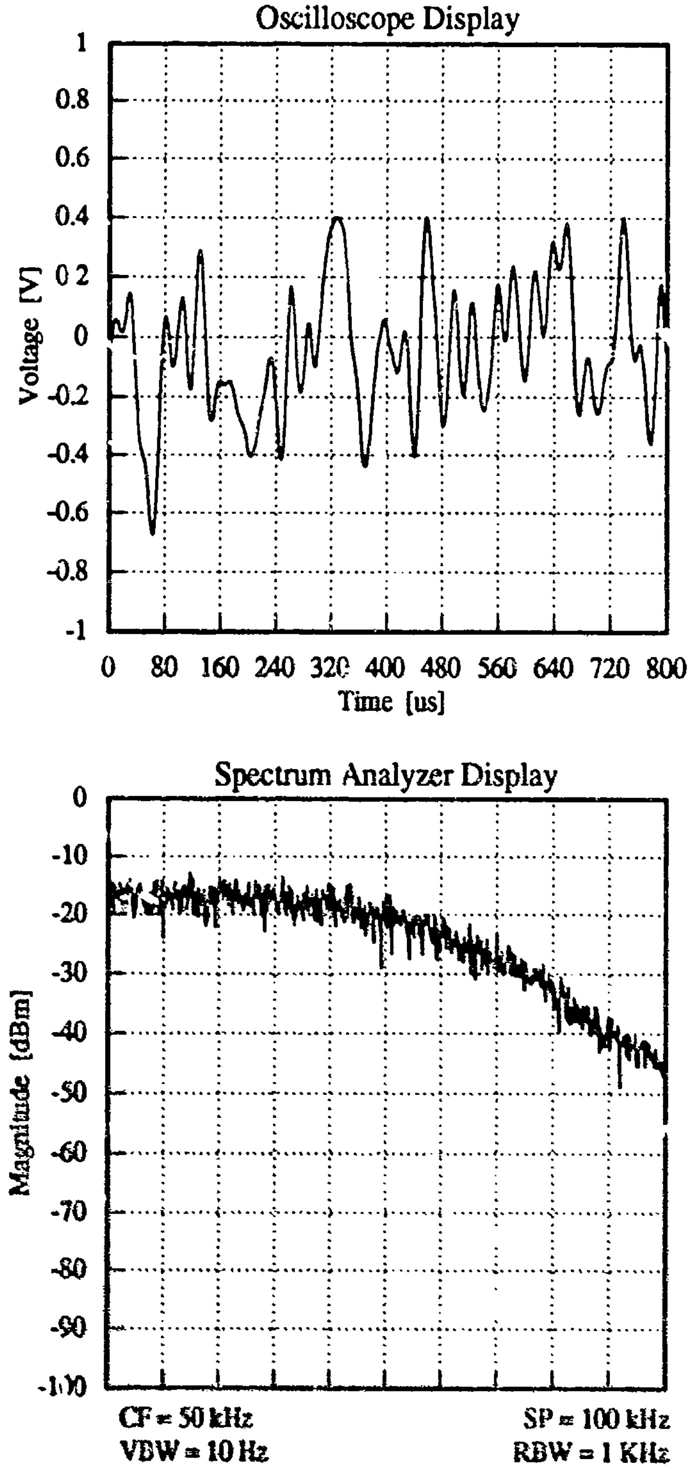


Figure 5-2 Typical Time and Frequency Domain Representations of 50 kHz Bandlimited White Gaussian Noise

ferent video bandwidths enabled the spectrum analyzer to record different features of the same FM-by-noise barrage. The Auto Display, positioned in the top half of each figure, shows a snapshot of the relatively instantaneous behavior of an FM-by-noise barrage . Positioned in the bottom half of each figure, the Manual Display illustrates the average behavior of the same barrage whose instantaneous behavior is illustrated in the Auto Display. The Auto Displays were measured with video and resolution bandwidths set to AUTO for the frequency span or span width selected. The Manual Displays, on the other hand, were recorded using a narrower video bandwidth than that used to record the Auto Displays. Hence, the traces acquired for the Manual Displays were averaged over time, while the traces acquired for the Auto Displays were relatively instantaneous.

It is important to note that the term *impulsive* in this thesis refers to frequency-domain impulses unless otherwise noted. Both the Auto and Manual Displays in Figures 5-3 through 5-5 illustrate the effect of increasing the bandwidth of the baseband noise while maintaining the peak frequency deviation constant. Note the impulsive nature—*frequency-domain impulsivity*—of the barrages decreases with increasing baseband noise bandwidth. That is, the barrages become increasingly fuller or continuous with increasing noise bandwidth. The Auto Displays illustrate the effect much more convincingly than the Manual Displays.

The impulsive nature of FM-by-noise in the frequency-domain can be related to one of the ratios mentioned in Chapter III. Specifically, the impulsive nature of FM-by-noise in the frequency-domain decreases as the deviation ratio, D, decreases. Recall, the deviation ratio, defined in Eqs. 3-8 and 3-9, is the ratio of peak frequency deviation to baseband signal bandwidth. The deviation ratio for Figure 5-3, Figure 5-4, and Figure 5-5 was 300, 30, and 3, respectively. That the impulsive nature of FM-by-noise decreases with D does not, however, imply that NBFM and UDRFM barrages produce more effective jamming than WBFM barrages. In fact, NBFM and UDRFM are, in general, not used for noise jamming because both NBFM and UDRFM do not sufficiently spread the baseband noise at RF.

Figures 5-3 through 5-5 also illustrate the description of FM-by-noise presented by Lothes *et al.* which was quoted in Chapter II (23:52). As alluded to in the quote and as explained in Chapter III, the instantaneous frequency of an FM-by-noise barrage is determined by its modulating noise via Eqs. 3-3

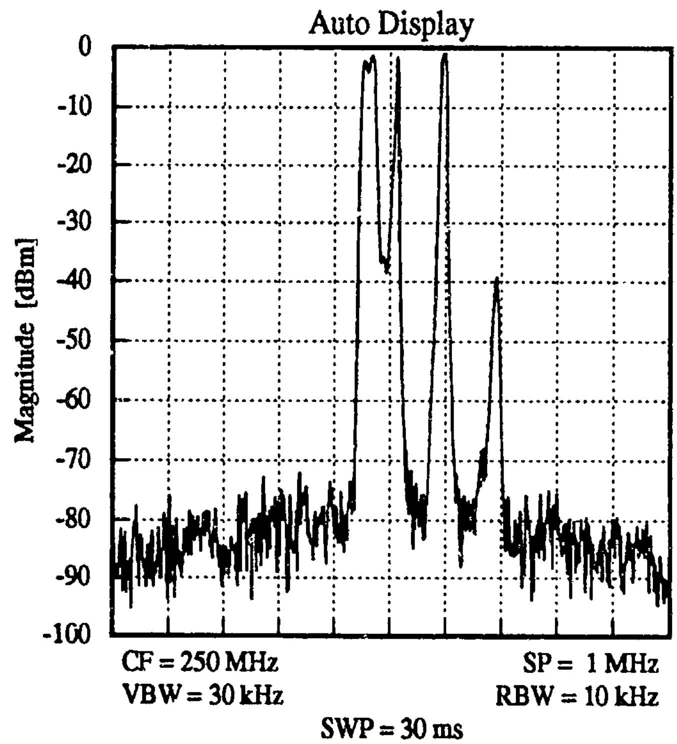
through 3-5. That is, the deviation of the baseband noise voltage about zero volts is a direct representation of the deviation of the jammer carrier about its center frequency. The barrage illustrated in Figure 5-3 (D = 300) was deviated more slowly about the jammer center frequency than the barrages illustrated in Figure 5-4 (D = 30) and Figure 5-5 (D = 3). However, the peak frequency deviation, Δf_p = 150 kHz, was the same in all cases. Of course, the theoretical barrage width, defined by Eq. 3-10, was the same for all the barrages.

Consistent with the theory on FM-by-noise, the jammer carrier modulated by 0.5 kHz noise took a longer time to visit each frequency contained in the full barrage width than the jammer carrier modulated by either 5 kHz noise or 50 kHz noise. Hence, the barrage modulated by 0.5 kHz noise, shown in Figure 5-3, is more impulsive in the frequency-domain than the barrages modulated by either 5 kHz or 50 kHz noise and shown in Figures 5-4 and 5-5 respectively.

Note that the theoretical bell-shaped spectrum associated with an RF carrier frequency-modulated by Gaussian noise and predicted by Woodward's Theorem is not immediately evident in either of the Auto Displays shown in Figures 5-3 and 5-4. It does become evident in the Auto Display of Figure 5-5 because of the decreased deviation ratio. All of the Manual Displays, however, exhibit the bell-shaped spectrum because, as mentioned earlier, the traces in the Manual Displays were recorded with a narrow video bandwidth which caused the traces to be averaged over time.

Also note that while the narrower video bandwidth caused the traces in the Manual Displays to exhibit the characteristic bell-shape indicative of a carrier frequency-modulated with Gaussian noise, it also averaged out the impulsive behavior of FM-by-noise evident in the Auto Displays. However, this lost information can be partially recovered by comparing the individual Manual Displays.

Consider, first, the barrage modulated by the 0.5 kHz noise depicted in Figure 5-3. Not only is the peak magnitude of its PSD less than peak magnitudes of the barrages modulated by 5 kHz and 50 kHz noise respectively, its trace line is also more thick, jagged, and ill-defined. Since the peak frequency deviation for all three barrages was equal, the magnitudes of each PSD and the uniformity of each trace can be used as a measure of how frequency-domain impulsive the actual barrages were. However, the degree of impulsivity can only be determined comparatively.



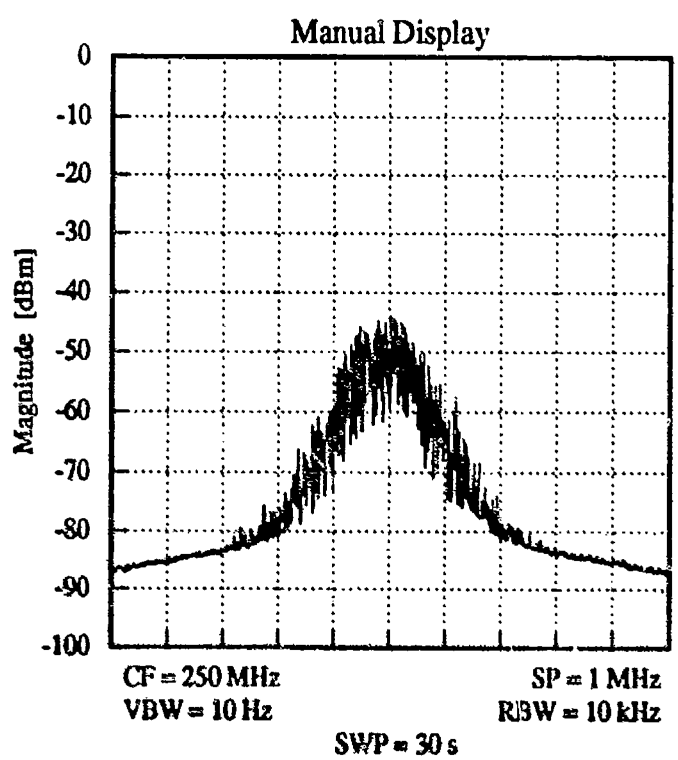


Figure 5-3 Spectrum Analyzer Displays of FM-by-Noise Barrage at RF: $\Delta f_{p} = 150$ kHz $B_N = 500$ Hz

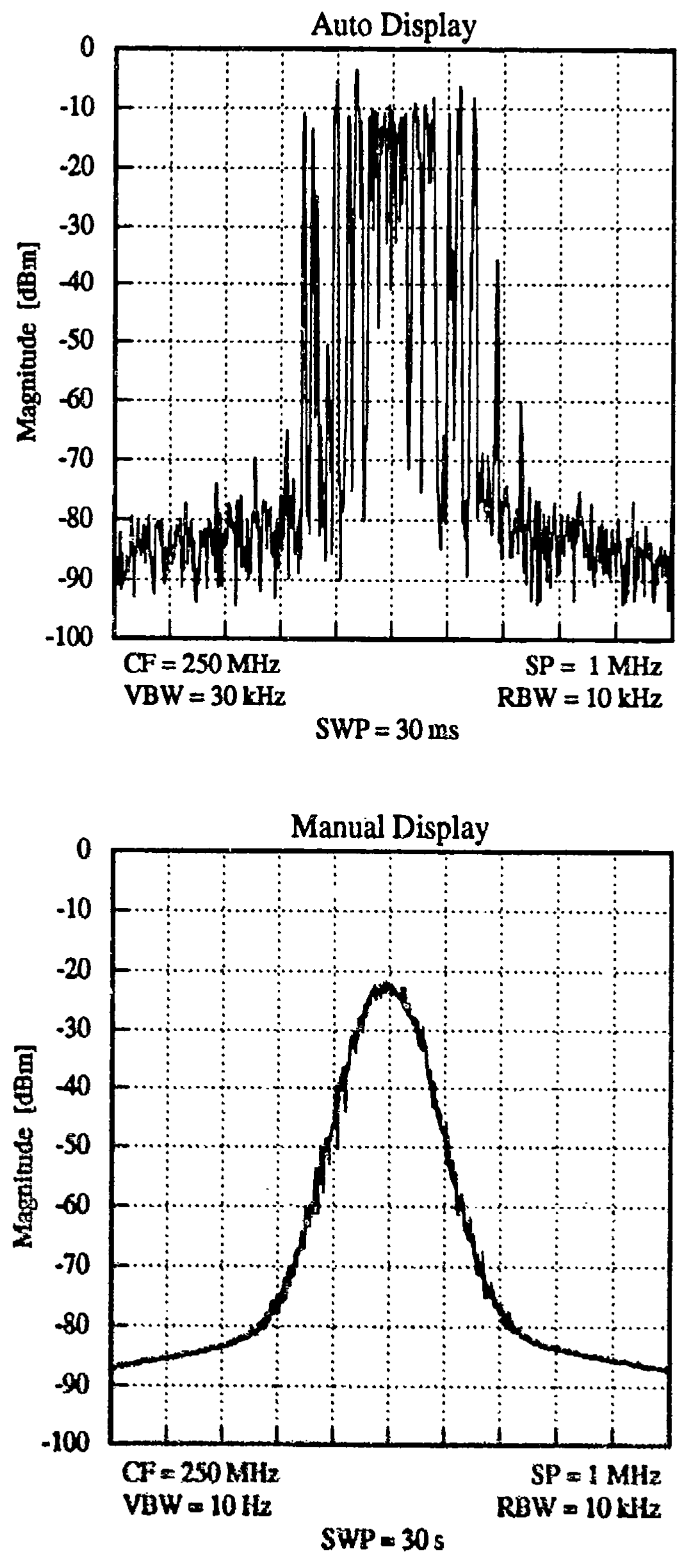
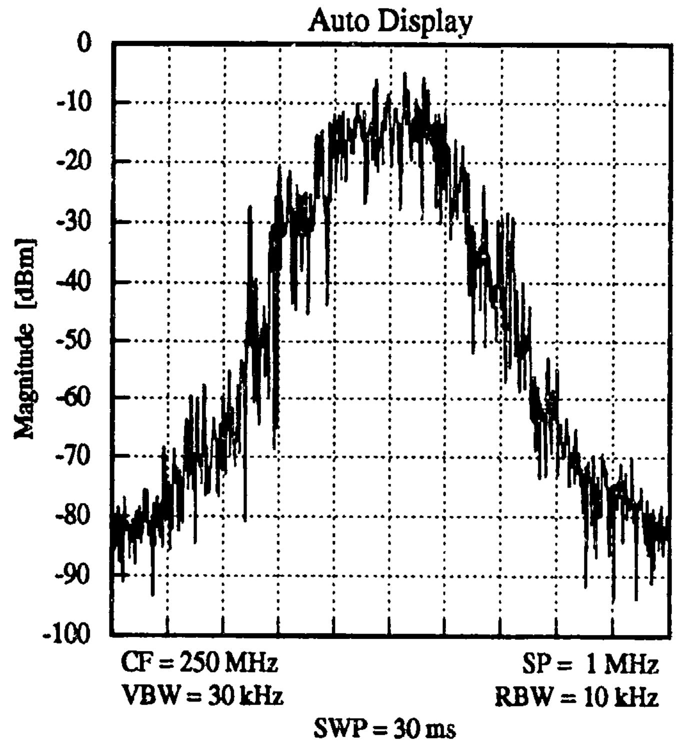


Figure 5-4 Spectrum Analyzer Displays of FM-by-Noise Barrage at RF: $\Delta f_p = 150$ kHz $\quad B_N = 5$ kHz



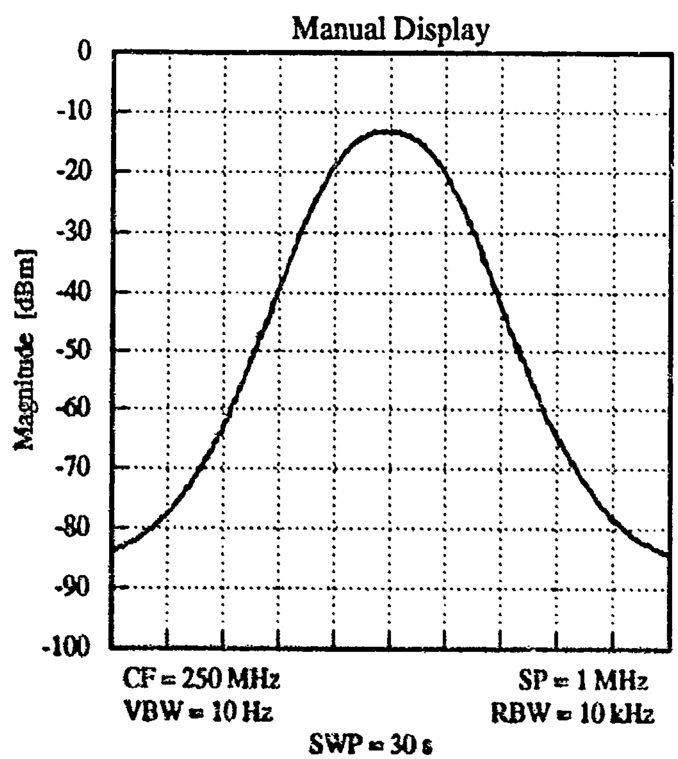


Figure 5-5 Spectrum Analyzer Displays of FM-by-Noise Barrage at RF: $\Delta f_p = 150$ kHz $\quad B_N = 50$ kHz

Given equal peak frequency deviations and equal video bandwidths, low-magnitude PSD barrages with thick, jagged trace lines indicate impulsive jamming. Hence, the barrage modulated by 0.5 kHz noise, shown in Figure 5-3, is more impulsive than the other two barrages shown in Figures 5-4 and 5-5. Note that the narrow video bandwidth used to record the Manual Displays irretrievably averaged out information on the individual impulses that would have otherwise been captured in the Auto Displays.

Figure 5-6 illustrates the well-documented power-bandwidth trade-off inherent in any noise jamming scenario. The center frequency was increased to 900 MHz so that a peak frequency deviation of 4 MHz could be selected. The frequency span or span width (SP) for the display at the top of Figure 5-6 is 1 MHz while the frequency span for the bottom display is 10 MHz. Note that the barrage with the lower deviation ratio ($D = 3$) had a higher peak PSD magnitude than the barrage with the higher deviation ratio ($D = 80$) but was not as wide. Thus, the power-bandwidth trade-off is demonstrated.

5.3 FM-by-Noise at IF

Figures 5-7 through 5-12 illustrate the effects of FM-by-noise at IF. All the figures have the same format. Channel One, the top part of the oscilloscope display, shows the baseband noise that was used to frequency-modulate the RF carrier which in all cases illustrated in Figures 5-7 through 5-12 was 250 MHz. The barrages shown in Figures 5-3 through 5-5 are representative of the jamming that was used to generate Figures 5-7 through 5-12. Channel Two, the bottom portion of the oscilloscope display, shows the time-domain output of the victim receiver's IF filter as the frequency-modulated jammer carrier was swept across its passband. The oscilloscope traces were recorded simultaneously using a sampling rate of 1 MHz and vertical resolution of 6 bits. As shown in Table B-2, the maximum resolvable frequency for these settings was 250 kHz. Channels One and Two were set for the same time/division; however, time rather than time/division is shown on the figures and it is annotated only once, with Channel Two, for readability.

The spectrum analyzer display shows the frequency-domain representation of the filter output. The displays are marked with the center frequency (CF) and trace span width (SP) in Hz. The filter bandwidth, peak frequency deviation, and baseband noise bandwidth associated with each FM-by-noise jamming scenario are explicitly stated in the figure captions. Video and resolution bandwidths were set to AUTO and

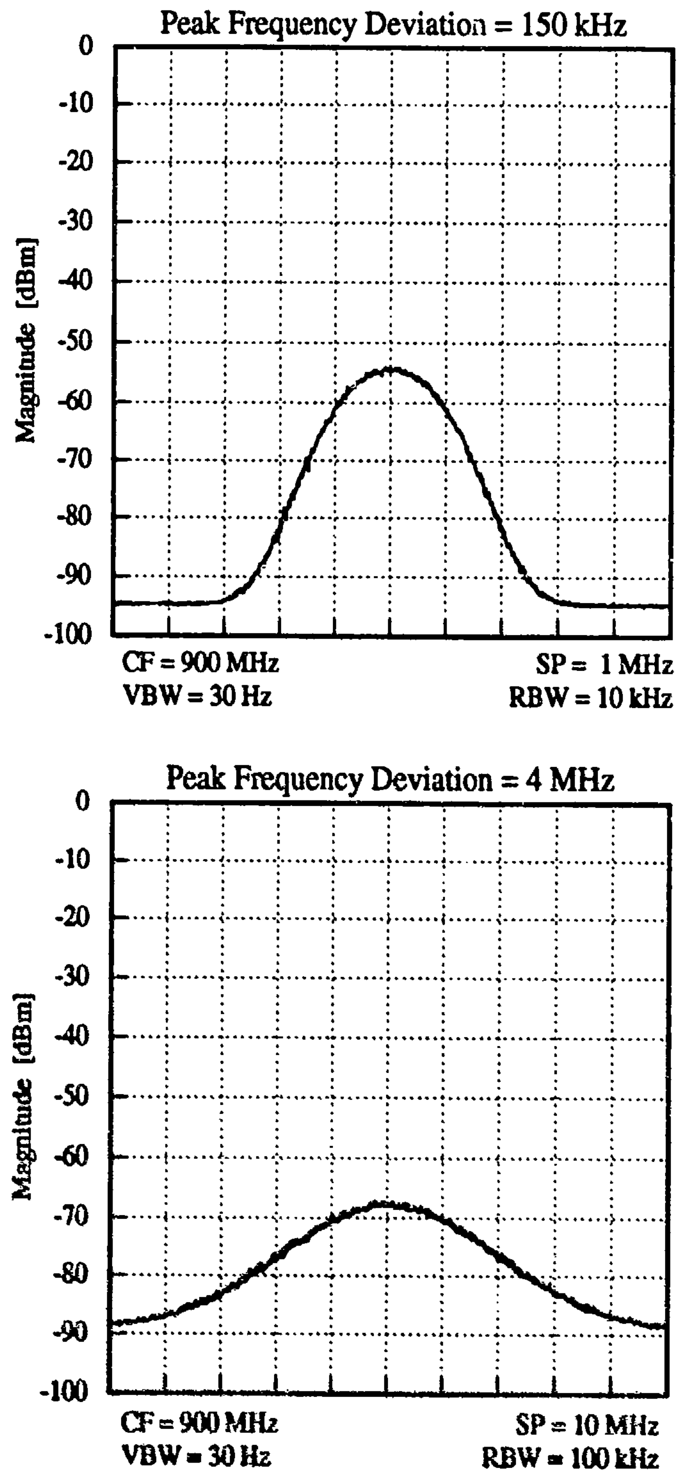


Figure 5-6 Spectrum Analyzer Displays of Two FM-by-Noise Barrages Modulated by 50 kHz Noise and Having Unequal Peak Frequency Deviations

are not included with spectrum analyzer displays but are mentioned here. The video bandwidth, resolution bandwidth, and sweep for the spectrum analyzer displays in Figures 5-7 through 5-12 were 3 kHz, 1kHz, and 300 msec, respectively.

It will prove beneficial to reiterate a few points on the theory of FM-by-noise before discussing the figures. Recall, the instantaneous frequency of an FM-by-noise barrage is determined by its modulating noise via Eqs. 3-3 through 3-5. That is, the deviation of the baseband noise voltage about zero volts is a direct representation of the deviation of the jammer carrier about its center frequency. Furthermore, if the jammer center frequency is tuned to the center frequency of the victim receiver, the deviation of the baseband noise voltage about zero volts is also a direct representation of the jammer carrier as it deviates about the center frequency of the victim receiver. The deviation of the jammer carrier about the center frequency of the victim receiver is also referred to as a sweep, or, as in (22:147), an excursion.

Recall also that the signal generator used in the experiments produced a peak deviation of $\pm\Delta f_p$ when the voltage at the FM input was nominally ±1 volt. Hence, the Voltage axis on Channel One of the oscilloscope displays may also be interpreted as a frequency axis. Since it is assumed that the RF center frequencies used in the experiments coincided with the center frequencies of the IF filters, a baseband noise voltage of zero volts corresponds to the situation where the jammer carrier exactly coincides with the center frequency of the victim receiver. As the baseband noise voltage deviates above or below zero volts, the FM-by-noise jammer deviates, sweeps, or makes an excursion across the victim receiver passband. In short, the voltages, $\pm V_{3dB}$, corresponding to the upper and lower edges of the 3 dB bandwidth of the IF filter are

$$\pm V_{3dB} = \frac{B_V}{2 \cdot \Delta f_p} \text{ Volts} \tag{5-1}$$

where a positive voltage corresponds to the upper 3 dB frequency of the filter and a negative voltage corresponds to the lower 3 dB frequency. Consider Figure 5-7. In light of the previous discussion, the voltage corresponding to the 3 dB bandwidth of the IF filter is, by Eq. 5-1, approximately ± 83 mv. As an aside, the term *passband* in these discussions refers to the bandwidth, swept by the jammer carrier, over which a discernible response can be detected on the Channel Two oscilloscope traces.

Focusing then on Channel One of the oscilloscope display in Figure 5-7, and, for the moment, ignoring any theory on FM-by-noise, it can be seen that the carrier started to sweep into the victim receiver passband with a positive slope at about the 320 μs mark and swept out some 320 μs later. Relying on Eqs. 3-3 and 3-4, it is suspected that the filter output will be a sinusoid of increasing frequency, due to the positive slope of the baseband voltage, that starts some short time after the 320 μs mark and lasts approximately 320 μs. Furthermore, relying on basic filter theory, it is suspected that the increasing-frequency sinusoid at the filter output will be amplitude-modulated with the shape of the filter transfer function because the various frequencies contained in the increasing-frequency sinusoid will be appropriately weighted by the filter transfer function. This is exactly the situation illustrated by the Channel Two display in Figure 5-7.

Figure 5-7 and the remaining figures in this chapter can be related to the theory, equations, and definitions developed in the references and the previous chapters. Figure 5-7 depicts an WBFM-LFN scenario in which the baseband noise voltage made a full sweep, with a positive slope, across the passband of the victim receiver. Consistent with the theory on FM-LFN, the filter output was a sine wave of increasing frequency which was amplitude-modulated with the shape of the filter. The filter output had a duty cycle corresponding to the length of time it took the carrier to sweep across the receiver passband. In short, the output was a pulse or ring that lasted approximately 320 μs. Note the term *pulse*, in this context, emphasizes the discrete nature of the output, while the term *ring* emphasizes the damped sinusoidal-nature of the output.

Since the sweep of the carrier across the passband was approximately linear, it can be compared with the variable, S, and the ratio, a, defined by Benninghof *et al.* (8:14-20–21) and described in Chapter III. The slope of the linear sweep from 320 μs to 640 μs is approximately 13 MHz/sec. This slope of 13 MHz/sec corresponds to the variable S. Next, the slope is divided by the 3 dB bandwidth of the victim receiver bandwidth squared, B_V^2. The quotient approximates the ratio a because the 3 dB bandwidth of the filter was used rather than b as defined in Eq. 3-24.

The result is approximately 0.021 and is qualitatively consistent with definition of a slow sweep, $a << 1$. Note that the SVR for the scenario being discussed is shown in Table 4-1 and is 0.12. While this

SVR does agree qualitatively with the ratio α, it does not agree quantitatively. Better quantitative agreement is obtained if an RMS DVR is used to calculate the SVR. Recall, the RMS DVR, for a Gaussian process, can be calculated from the DVR by simply dividing by three. Considering an RMS DVR then, the SVR is 0.04 .

Next, consider the spectrum analyzer display of the filter output shown in Figure 5-7. The jamming is not continuous or full. This result is not surprising when one considers the previous discussion of FM-by-noise at RF and examines the barrage shown in Figure 5-3 used to produce the jamming. Specifically, the barrage was impulsive in the frequency-domain at RF. Hence, that the resulting jamming at IF is also impulsive is not surprising. In general, WBFM-LFN jamming is frequency-domain impulsive in nature.

The jammer center frequency, as it relates to Figure 5-7, does not coincide exactly with the victim receiver center frequency, but it is close enough that zero volts on Channel One may be considered to correspond to the victim receiver center frequency. This will be the case throughout the chapter; however, the mismatch between the jammer center frequency and victim receiver center frequency will be mentioned only when necessary.

Extrapolation to the more real-world case when the jammer center frequency is off victim center is straightforward. In this case, some non-zero value of the baseband noise voltage will correspond to the center frequency of the victim receiver, and the filter output will be a maximum as the baseband sweeps through this non-zero value. The exact value of the baseband noise voltage that corresponds to the center frequency of the victim receiver will depend on the sign and magnitude of the mismatch between the jammer center frequency and the center frequency of the victim receiver.

Like Figure 5-7, Figure 5-8 also illustrates a WBFM-LFN jamming scenario. The time-domain output of the filter shown on Channel Two of the oscilloscope consists of five pulses or rings. The first ring on the left was caused by the jammer as it just barely entered the receiver passband and then turned back out. Hence, this first ring has a small amplitude. The next pulse was caused as the jammer made a full sweep across the receiver passband. Note the maximum amplitude of the pulse occurs some time after the baseband voltage sweeps past zero volts. This observation is due to filter delay and jammer-victim frequen

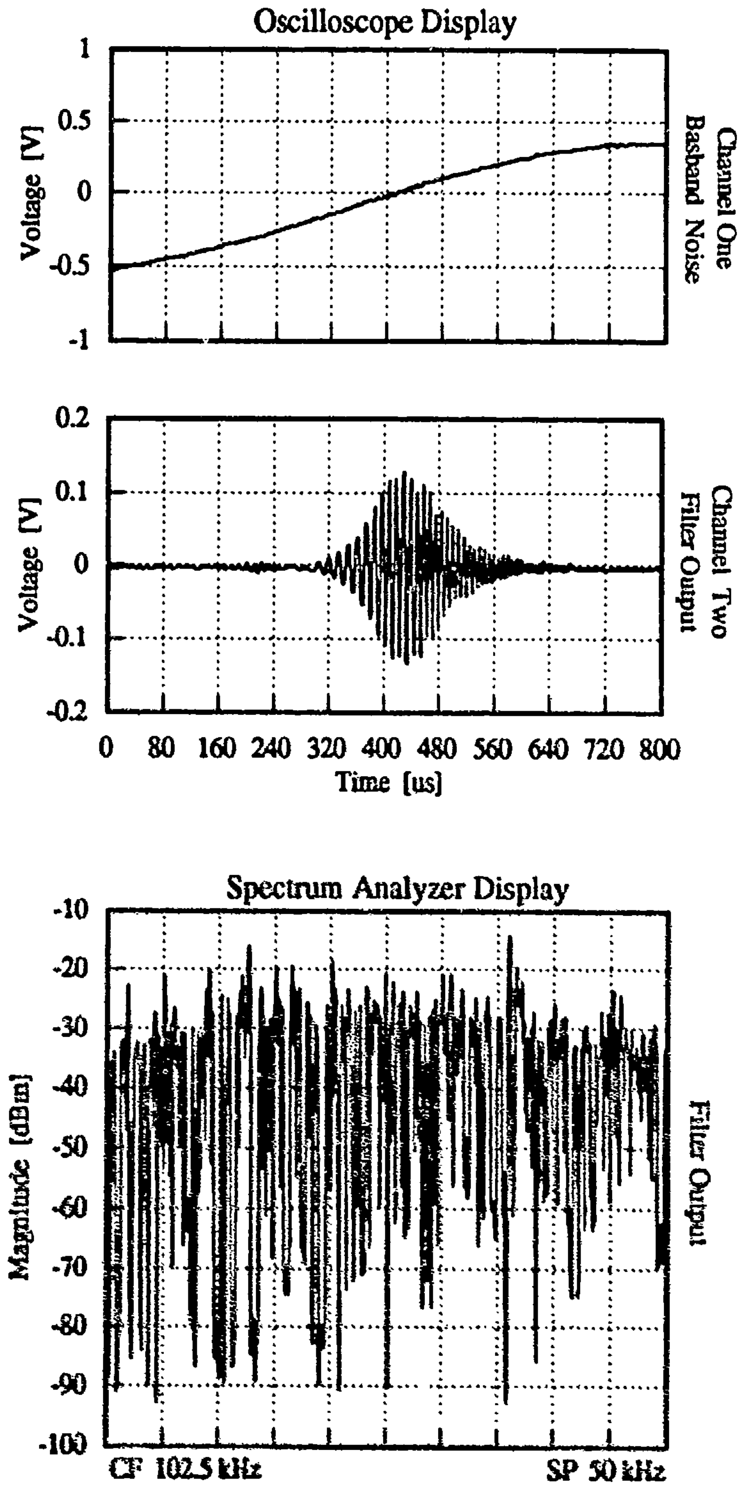


Figure 5-7 Response of 25 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $\quad B_N = 0.5$ kHz

cy mismatch. The jammer remained outside the receiver passband for approximately 160 μs and then veered back into the passband. Rather than making a full sweep, the jammer lingered near the center frequency of the filter for approximately 80 μs. Hence, the amplitude of the ringing is relatively high, its frequency content is relatively constant, and its duty cycle is approximately 80 μs. The jammer turned back out of the receiver passband but returned later for two more full sweeps.

The slope of the fourth sweep is steep, so the output pulse due to this steep sweep is impulsive in the time-domain. In light of the previous discussion, the relation between the baseband noise and the final pulse is self-explanatory. Note that the first and third rings illustrate the effect of FM-LFN when the jammer carrier does not make a full sweep across the receiver passband. Consistent with the theory, the rings have amplitudes appropriately weighted by the filter transfer function and duty cycles corresponding to the amount of time the jammer was actually in the receiver passband.

Finally, the spectrum analyzer display in Figure 5-8 shows a filter output that is less frequency-domain impulsive than that shown in Figure 5-7. This result is not surprising when one compares the frequency-domain structures of the FM-by-noise barrages, shown in Figures 5-3 and 5-4, used to cause the jamming. Clearly, the more frequency-domain impulsive the barrage is at RF, the more frequency-domain impulsive the resulting jamming will be at IF. The time-domain structure of the baseband noise processes shown in Figures 5-7 and 5-8 also help explain why the resulting jamming in the latter scenario is less frequency-domain impulsive. Since the 0.5 kHz baseband noise used to generate the jamming shown in Figure 5-7 had, and has by definition, a lower frequency content than the 5 kHz baseband noise shown in Figure 5-8, the 0.5 kHz baseband noise was guaranteed to have less zero-crossings than the 5 kHz baseband noise. Less zero-crossings translate to less sweeps across the receiver passband, and less sweeps across the receiver passband translate to less pulses or rings at the filter output. Hence, the time-domain structure of the baseband noise coupled with the peak frequency deviation and victim receiver bandwidth give an indication of how impulsive, in the frequency-domain, the resulting jamming will be.

As was done for the jamming scenario illustrated in Figure 5-7, the sweep S and the ratio a can be calculated for each full passband excursion observed in Figure 5-8. The S and a associated with each full

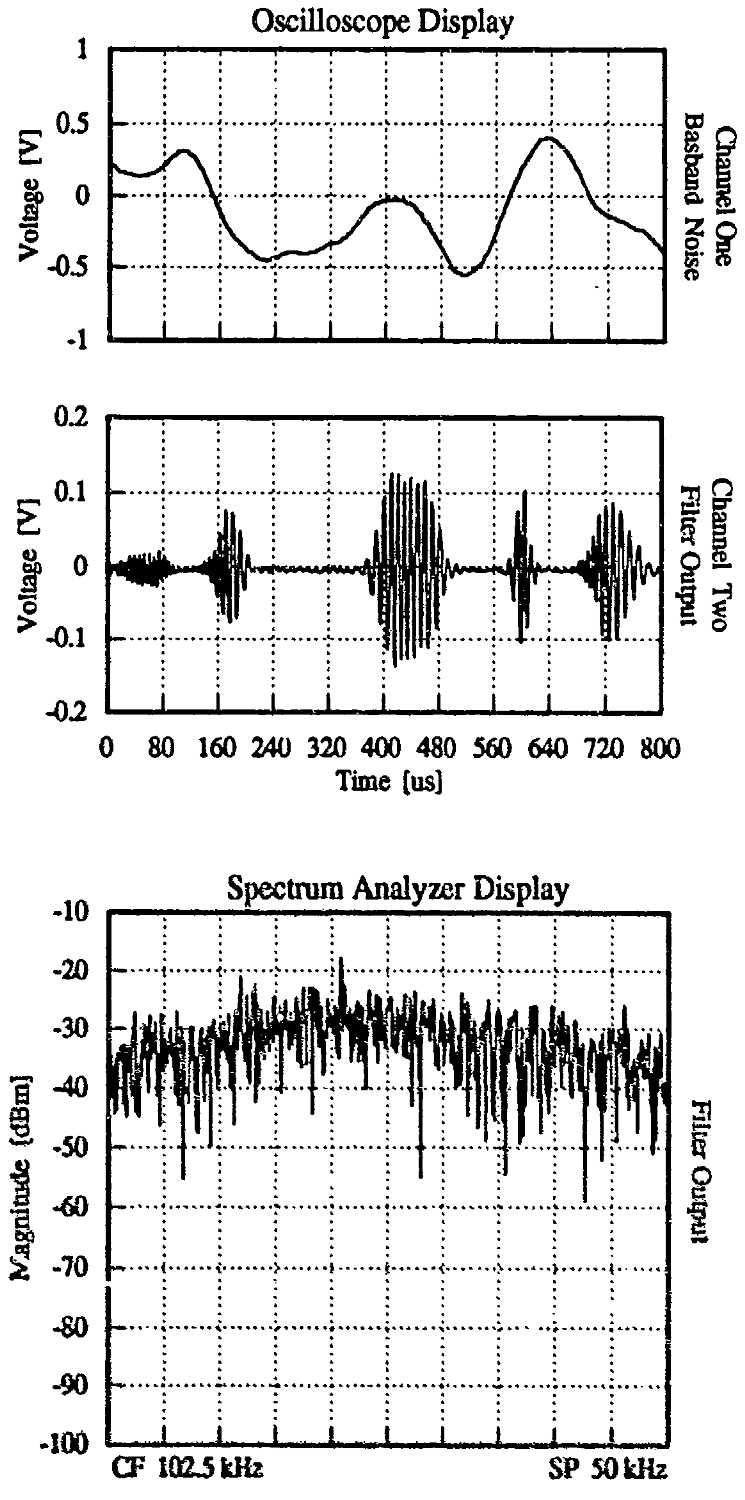


Figure 5-8 Response of 25 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $B_N = 5$ kHz

passband excursion could then be compared with the SVR for this scenario. However, these routine calculations and comparisons will not be made since no new or profound insights into the nature of FM-by-noise will be gained by doing so. Instead, the SVR and its relation to the filter output will be discussed qualitatively. The SVR, 1.2, is listed in Table 4-1. This SVR is more close to the definition of a slow sweep, $a \ll 1$ than to the definition of a fast sweep, $a \gg 1$, and the filter output is, as predicted by Benninghof *et al.*, "...a series of impulses [or pulses or rings] of random amplitude and random structure" (8:14-20–21).

Figure 5-9 illustrates a WBFM-WBN jamming scenario. Unlike the previous two examples, however, the baseband noise caused the jammer to deviate about the receiver center frequency much more rapidly and more often as evidenced by the number of zero-crossings the baseband noise made. Hence, it is somewhat difficult to relate the baseband noise and the filter output in the time-domain. Note that the SVR for this case is, as shown in Table 4-1, is 12 which corresponds to the definition of a fast sweep, $a \gg 1$. Consistent with the description of FM-WBN, the number of full sweeps across the victim receiver passband has increased in comparison to the previous two examples. For the most part, the filter does not have time to recover from one sweep before another full sweep occurs. Consequently, the ringing responses overlap, and the output is noise. As evidenced by the Channel Two trace shown in Figure 5-9, the filter appears to have had time to recover from the overlapping rings at least twice. Note that the baseband noise strayed away from zero volts for an appreciable length of time between the 160 μs and 240 μs mark and again around the 480 μs mark. Not surprisingly, the filter output became negligibly small during these times.

According to the theory on FM-by-noise, the output of a filter subjected to FM-WBN should be approximately Gaussian. That this is the case is not immediately evident from Figure 5-9. Measuring the degree of normality, or Gaussianity, of jamming is the topic covered in the next chapter where the concept of noise quality is reexamined. Finally, the jamming at the output of the filter seems to be quite full as indicated by the spectrum analyzer display. Again, this frequency-domain result is not surprising when one considers not only the time-domain structure of the baseband noise but also the frequency-domain structure of the FM-by-noise barrage, shown in Figure 5-5, used to produce the jamming.

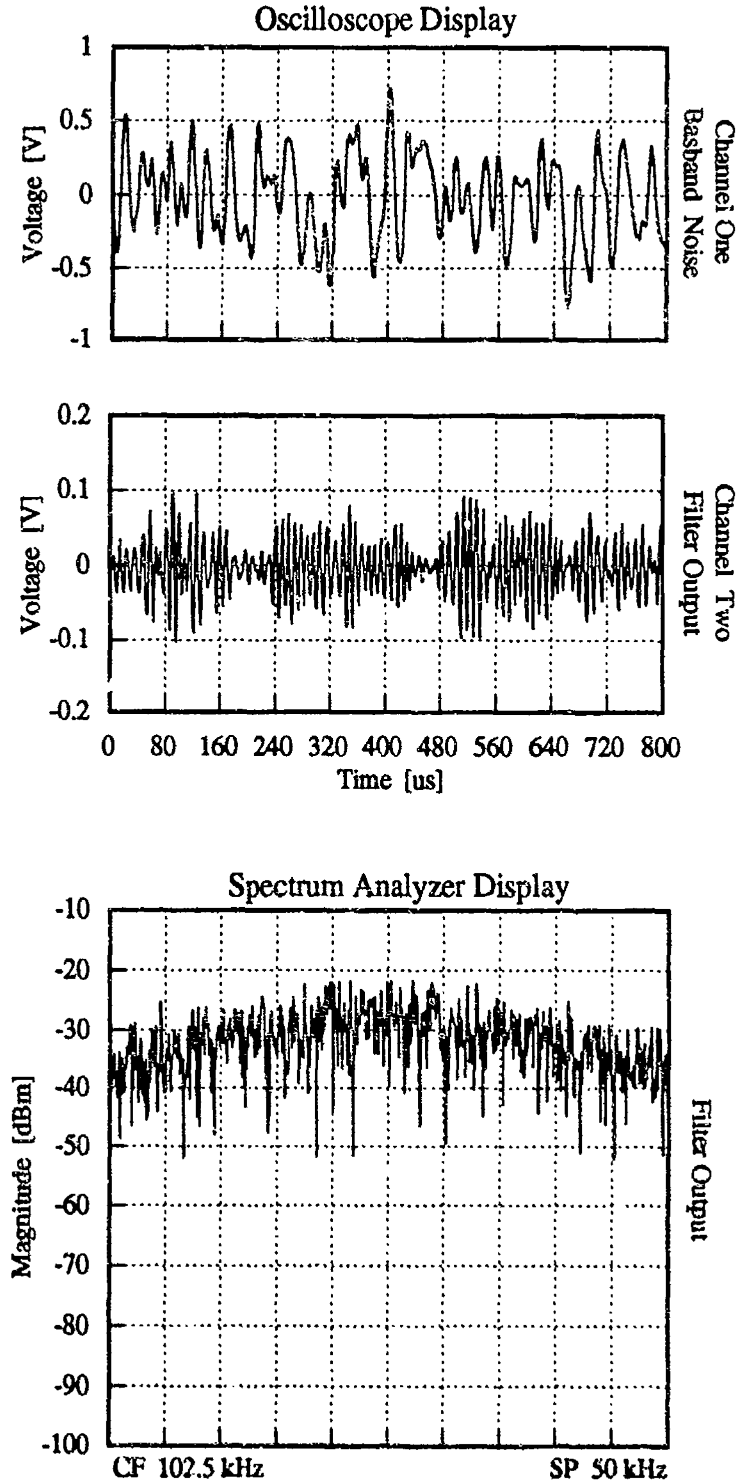


Figure 5-9 Response of 25 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $\quad B_N = 50$ kHz

Figures 5-10 through 5-12 exhibit features very similar to the FM-by-noise jamming scenarios illustrated in Figures 5-7 through 5-9 which have already been discussed in detail. Hence, lengthy discussions of Figures 5-10 through 5-12 are not necessary. Instead, general comments will be made. The only difference between the scenarios shown in Figures 5-7 through 5-9 and Figures 5-10 through 5-12 is the victim receiver bandwidth. The victim receiver bandwidth used in the latter set of figures was increased from 25 kHz to 50 kHz. Therefore, Figures 5-10 and 5-11 illustrate FM-LFN scenarios while Figure 5-12 illustrates a FM-UBN scenario. FM-UBN is, for the purposes of this investigation, similar to FM-WBN.

5.4 Summary

To complement the theory on FM-by-noise, this chapter presented results of FM-by-noise jamming experiments conducted using commercial test and measurement equipment. Specifically, the time- and frequency-domain behavior of FM-by-noise jamming at RF and IF, in terms of the DVR, NVR, and SVR, was studied with the baseband noise bandwidth, peak frequency deviation, and victim receiver bandwidth as parameters. An important feature of this chapter was the inclusion of computer-generated reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise at RF and IF. The behavior of FM-by-noise illustrated in the oscilloscope and spectrum analyzer displays was related back to the description of FM-by-noise jamming developed in the analytical portion of the investigation.

The RF frequency-domain structure of an FM-by-noise barrage modulated by white Gaussian noise was found to be related to the baseband noise bandwidth and peak frequency deviation. For a given peak frequency deviation, FM-by-noise barrages modulated by wideband noise are less frequency-domain impulsive than FM-by-noise barrages modulated by narrowband noise. The behavior of FM-by-noise at IF was shown to be dependent on the frequency-domain structure of the corresponding FM-by-noise barrage at RF and the bandwidth of the victim receiver. FM-LFN was shown to produce discrete rings or pulses at the output of the IF filter of the victim receiver, while FM-UBN and FM-WBN approximated the effects of DINA.

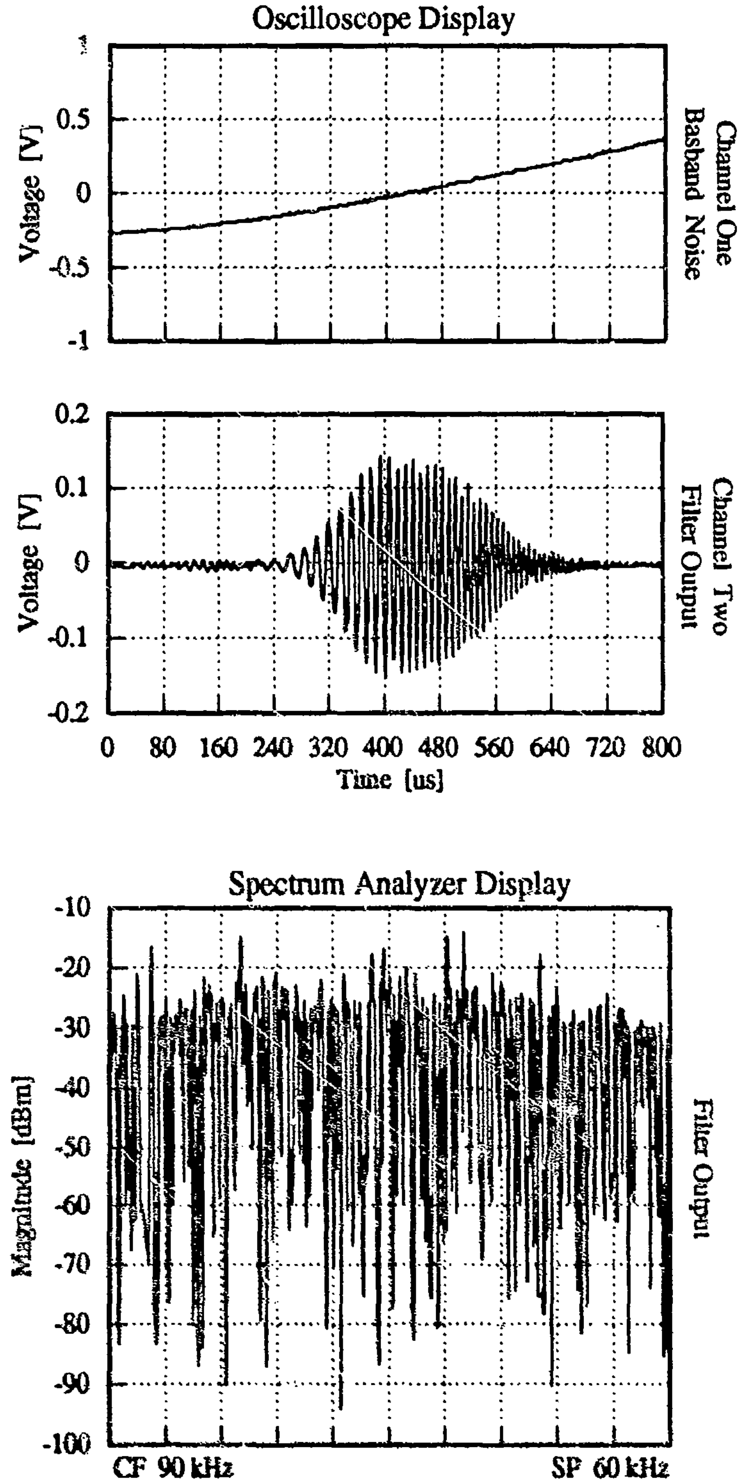


Figure 5-10 Response of 50 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $\qquad B_N = 0.5$ kHz

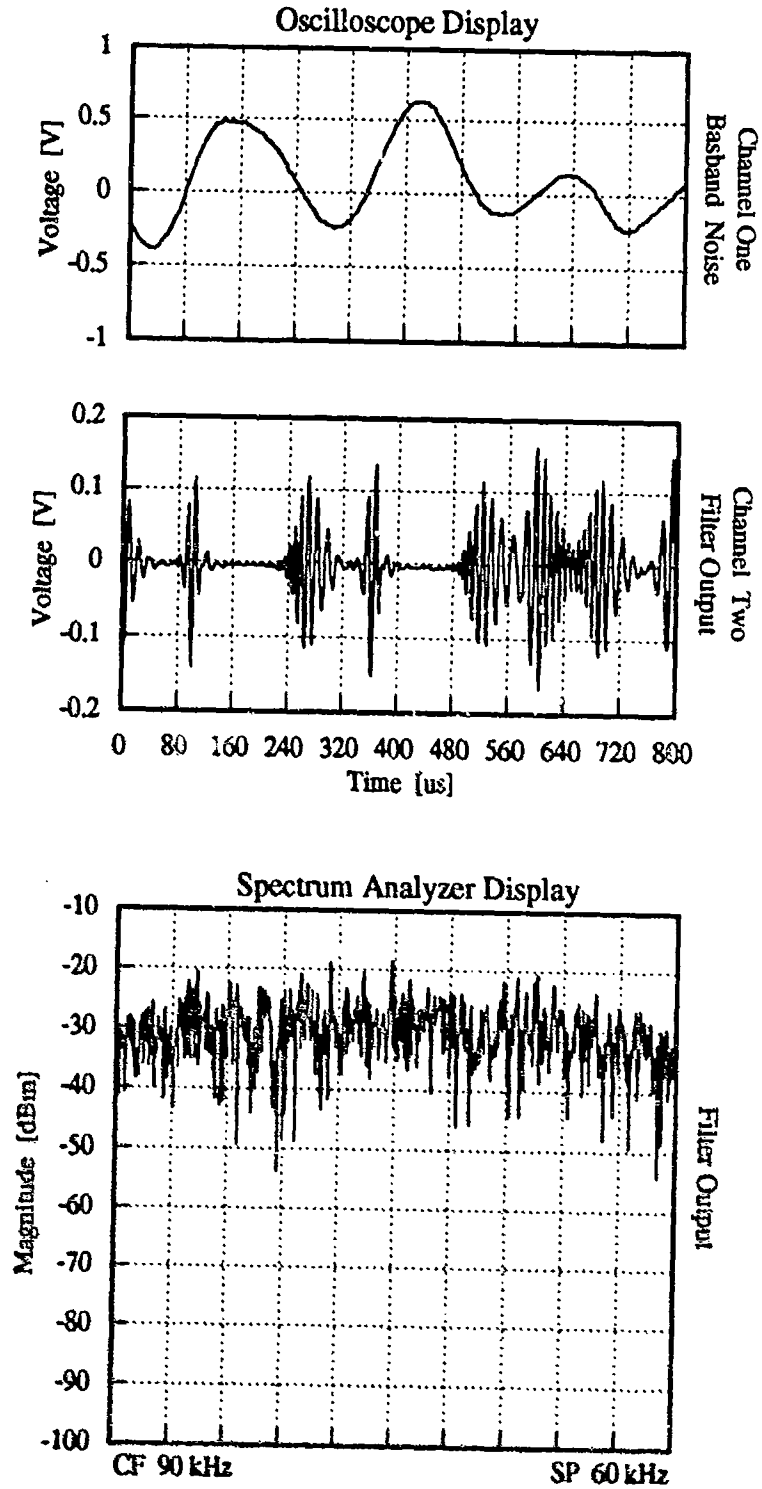
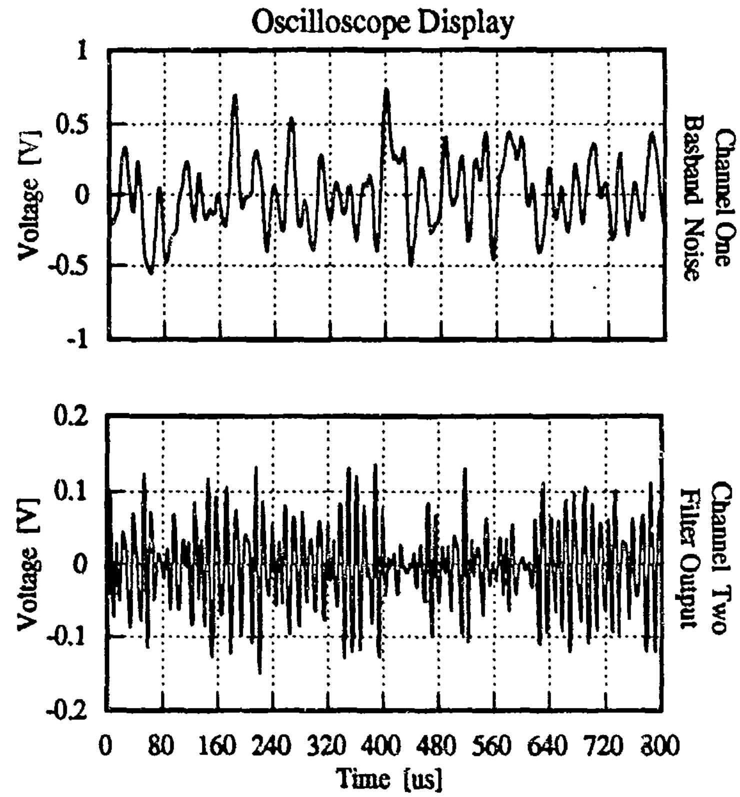


Figure 5-11 Response of 50 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $\quad B_N = 5$ kHz



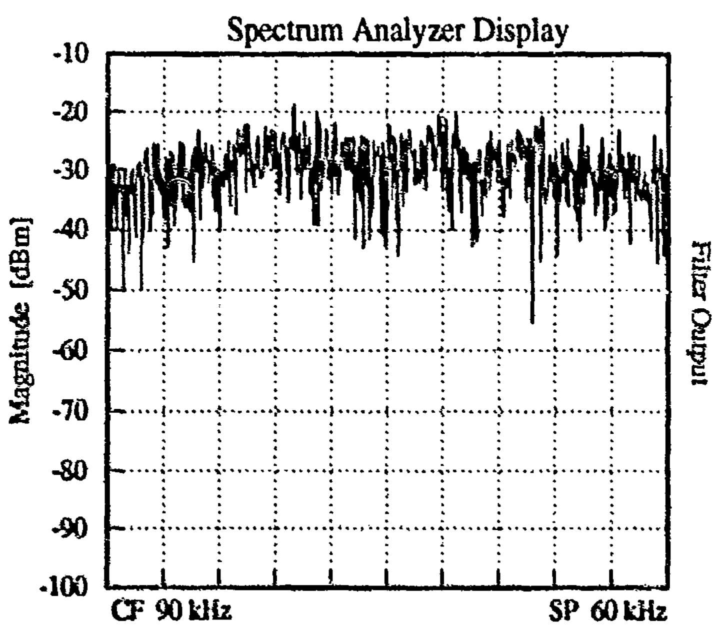


Figure 5-12 Response of 50 kHz Filter to FM-by-Noise:
$\Delta f_p = 150$ kHz $B_N = 50$ kHz

VI. Noise Quality Revisited

The material presented thus far deliberately avoided any explicit, quantitative discussion on jammer effectiveness but, instead, alluded to jammer effectiveness in strictly qualitative terms. In the late 1960's, the United States Air Force contracted with Stanford Electronic Laboratories to research methods of measuring noise jammer effectiveness. Researchers at Stanford investigated and reported on four tests to measure jammer effectiveness (29:40). Turner *et al.* extended the Stanford team's work and demonstrated that the effectiveness of a given noise jammer could be described by a single, albeit *ad hoc*, quantity which they called *noise quality* (43). In this chapter, the work done by Turner *et al.* is revisited and two alternative noise quality measures are proposed. A review of the literature on noise quality is presented first. It must be emphasized that noise quality is used to measure the effectiveness of noise jamming and is, in most cases, not applicable to deception jamming.

The three noise quality measures were programmed and preliminary testing was conducted. Time constraints prevented a thorough review and rigorous statistical analysis of the data obtained. The raw data obtained from these preliminary tests are included in Appendix C. Trends evident in the raw data are commented upon, in a qualitative manner, in this chapter. Equipment settings referred to in this chapter are capitalized; see Chapter IV for procedures on equipment set-up.

6.1 Literature Review

Shannon laid the foundations for noise quality in his celebrated works on information theory and communication theory (36–38). Shannon proved that the information entropy of a one-dimensional, continuous density of a given mean square value is maximum when the density is Gaussian (37:21). He also defined a quantity he called *equivocation*, or conditional entropy, to describe the amount of information lost when a message is transmitted over a noisy channel (36:13). Using these facts, he later proved that "...white Gaussian noise is the worst among all possible noises..." in terms of channel capacity and communications reliability (38:40). Hence, noise jamming which produces white Gaussian noise in a victim receiver should be the best jamming possible in most cases where conventional coding, modulation, and de-

modulation techniques are used. Noise quality attempts to quantify how good a given noise jammer is at producing white Gaussian noise in the IF filter of a victim receiver.

Numerous authors followed Shannon and either extended or expounded on his seminal work. Just a few of these authors are mentioned here. The compactness of Shannon's original works prevented an adequate discussion of the mathematical subtleties of his theories. Both Ash and Gallager discuss the mathematical subtleties of Shannon's theory at great length. Additionally, they present cogent descriptions of Gaussian channels. (2:230–261; 17:355–441). In a more practical vein, Beckmann discusses channel capacity, continuous systems, and applications of Shannon's channel-capacity formula (6:361–401). Finally, Lathi's text contains a readable introduction to information theory (21:428–479). These references provide material essential to the development of noise quality, but they do not specifically address the issue of noise quality.

Texts on spectral analysis, time-series analysis, and applied statistical analysis are indispensable to the development of noise quality. Priestley presents a complete and scholarly discourse on spectral and time-series analysis in (30). In fact, he mentions various goodness-of-fit tests, including tests for white noise (30:475–494). A thorough comparison of goodness-of-fit techniques is the sole topic of the text edited by D'Agostino and Stephens (15). Readable texts on time-series analysis from an engineering point of view are (7) and (28). In his text on applied statistical signal analysis, Shiavi presents numerous applications and examples including a white-noise test based on autocorrelation analysis (39:199).

Although noise quality is mentioned explicitly in a few radar and EW texts, it is usually left mathematically undefined (3:140; 4:12-5–12-6; 23:52; 33:129). Schleher, writing in (33), refers the reader to the work of Turner *et al.* (43). Maskimov *et al.* provide a mathematical definition of noise quality that is intuitively appealing, but their application of the measure to FM-by-noise jamming is abstruse (24:33–43).

As already mentioned, the United States Air Force commissioned a study on methods of measuring noise jammer effectiveness in the late 1960's (29), and an outgrowth of (29) was the *ad hoc* definition of noise quality proposed by Turner *et al.* in 1977 (43). The definition of noise quality proposed by Turner *et al.*, referred to as Turner noise quality in the sequel, will be discussed more fully in the next section. Fi-

nally, Knorr and Karantanas present results obtained from a computer simulation of Turner noise quality done in 1985 (20).

6.2 Turner Noise Quality

Since Turner noise quality (43) is based on research done at Stanford in the late 1960's (29), it will be beneficial to review the highlights of that research. The team at Stanford was tasked with developing methods of measuring the effectiveness of noise jammers. In light of Shannon's work, they hypothesized that the effectiveness of a given noise jammer could be determined by measuring the first order probability density of the output of a receiver subjected to jamming from the noise jammer under test and, subsequently, comparing this measured density to an ideal normal density having the same mean and variance. In short, they jammed a receiver and measured the normality of the time-series at the output of the front end or IF filter of the receiver. They did not directly measure the uniformity of the PSD at the output of the filter.

At the time, measuring the first order probability density of a time-series was no small task. The Stanford team engineered an equipment set-up specifically designed for measuring the first order probability density of a time-series and, subsequently, comparing the measured density with an ideal normal density having the same mean and variance. Four tests were used to measure normality.

The first test was based on error measures. Three quantities — average error, e_a; rms error, e_r; and summed error, e_s — measured the numerical difference between the measured and ideal densities. The next test was a locational one which compared the kurtosis, k, and skewness, s, of the measured density to the known kurtosis and skewness of the normal density. The third test, based on information theory developed by Shannon, used relative entropy and the entropy power ratio as an indication of normality. Finally, the researchers used three statistical hypothesis tests, one of which was the chi-square goodness-of-fit test, to gauge normality (29:40).

The Stanford team reported that the first three tests could be used to measure jammer effectiveness but questioned the usefulness of statistical hypothesis tests, the fourth method of gauging normality, in measuring jammer effectiveness. They concluded their report by noting that the "...system designer must determine how much gaussianity [normality] is required for a given application. He must decide which of

the gaussianity measures he wishes to use and then set an acceptable upper bound" (29:41). Again, the Stanford team did not directly measure the uniformity of the PSD at the output of the filter.

Turner *et al.* combined the first three tests and defined noise quality as

$$\frac{1}{\text{Noise Quality}} \triangleq \frac{1}{3}\cdot\left[\frac{e_a+e_r+e_t}{3}+\frac{|k-3|+s}{2}+|\text{relative entropy in bits}|\right] \qquad (6\text{-}1)$$

This definition of noise quality was reportedly used to measure the effectiveness of several operational jammers with great success (43). Over the years, the effectiveness of a whole generation of noise jammers was evaluated, but operational requirements forced the dismantling of the equipment used to measure Turner noise quality. A new generation of noise jammers has been entering the Air Force inventory, and a means of independently evaluating their effectiveness is needed. Hence, an interest in noise quality has been rekindled.

6.3 Turner Noise Quality Revisited

The sponsors of this investigation requested that the work of Turner *et al.* (43) be reproduced. Appendix A contains the program listing for Turner noise quality as implemented for this investigation. The program is based on algorithms presented in (29) and Eq. 6-1 (43). A chi-square goodness-of-fit test, which does not appear in the original algorithm (43), was added. Note the chi-square test added to the Turner noise quality algorithm is conducted at the 0.05 level of significance and includes Yate's correction (13:562-567). The equipment set-up is described in Section 4.4.3 of Chapter IV

The components of Turner noise quality given in Eq. 6-1 were computed, by the Stanford team, using 5 million samples (29:43). In this investigation, however, long data-acquisition times and extensive data-processing requirements limited the number of samples that could be used to measure Turner noise quality. Time-series data at the output of the filter being jammed was acquired from the HP 54111D digitizing oscilloscope in data blocks containing 8192 samples. That is, the oscilloscope digitized the signal being measured and stored 8192 consecutive samples in memory. The digitized samples were subsequently downloaded from oscilloscope memory to the computer. The memory depth, in terms of time, was determined by the sampling rate. See Table B-2 in Appendix B. With vertical RESOLUTION filters set to OFF for fast

data-acquisition times, downloading one memory block or 8192 samples from HP 54111D memory to the computer still took approximately twenty seconds. At this rate, it would have taken approximately 3.4 hours to acquire 5 million samples. Adding to the data-acquisition time was the time required for data-processing once the data was read into the computer. Specifically, the data had to be decorrelated and converted from digital values to real values before any statistical analysis could be performed.

Recall independent samples are required to estimate the pdf of a stationary random process (39:192), and Turner noise quality compares the measured pdf of a signal to an ideal normal density with the same mean and variance. Also recall that although stationarity and ergodicity were assumed at the start of this investigation, independence of samples was not. Indeed, the samples acquired by the HP 54111D digitizing oscilloscope were correlated because of the nature of the signals being measured coupled with the sampling rates used to digitize the signals. More specifically, the data were correlated because the signals measured in this investigation were bandpass signals sampled at a lowpass rate. Consequently, all samples downloaded to the computer could not be used to estimate the pdf of the signal being measured. The number of correlated samples contained in each block of 8192 samples acquired by the oscilloscope depended on the bandwidth of the IF filter and the time interval between samples which, in turn, depended on the sampling rate selected on the HP 54111D oscilloscope. An example will help explain.

Consider the first FM-by-noise jamming scenario described in Table 4-1 and investigated in the experiments. All values in this example are approximate. The IF filter in the realization of this scenario had a bandwidth of 25 kHz and a center frequency of 102.5 kHz. The upper cut-off frequency of the filter was 115 kHz and its *noise correlation duration* was 40 μs (7:140–144). The noise correlation duration gives an indication of the time interval required between samples such that the samples are uncorrelated. Since the autocorrelation function for bandpass, white noise, $R_{xx}(\tau)$,

$$R_{xx}(\tau) = aB_V\left(\frac{\sin \pi B_V \tau}{\pi B_V \tau}\right)\cos 2\pi f_c \tau \qquad (6\text{-}2)$$

is zero at $\tau = 1/B_V$, the noise correlation duration was assumed to be $1/B_V$ (7:114). This assumption can also be made when lowpass, bandlimited white noise is being measured.

Because the upper edge of the filter was 115 kHz and because the output of the filter could not be bandpass-sampled without substantial pre- or post-processing, a sampling rate of 500 kilosamples/sec was selected on the HP 54111D oscilloscope. Sampling the output of the filter at a lower rate would have caused aliasing. See Table B-2 and its explanation in Appendix B. However, sampling at 500 kilosamples/sec caused the time interval between samples to be approximately 2 μs. This time interval was much less than the 40 μs noise correlation duration of the 25 kHz IF filter. Hence, the samples were correlated, and there were approximately 20 redundant or correlated samples between each pair of uncorrelated samples. In essence, because the output of the filter was a bandpass signal sampled at a lowpass rate, it was oversampled. As a result, the samples, in this example, were correlated (7:337).

Indeed, all Turner noise quality measurements in this investigation involved bandpass signals being sampled at a lowpass rate. Hence, all samples acquired for the Turner noise quality measurements were correlated. All correlated samples download from the oscilloscope were ignored, and the data was decorrelated by creating an array of uncorrelated samples. The uncorrelated values were converted from digital levels to real values before creating the array of uncorrelated samples. See lines 540 through 1250 of the Turner noise quality program in Appendix A.

After the voltage samples were acquired from the HP 54111D oscilloscope, converted to real values, and decorrelated, the sample mean and variance were calculated. Next, the uncorrelated samples were normalized for a zero mean and unity variance. See lines 1270 through 1450. Following Bendat and Piersol, a bin width of $0.2 \cdot \sigma$ was assumed, where σ is the measured standard deviation (7; 269). Since this assumption can, in general, produce a non-integer number of bins, the number of equal-width bins, K, was forced to be an integer. The bin width was recalculated based on this integer number of equal-width bins. This precaution ensured that there were no gaps or overlaps between bins. The samples were subsequently sorted into $K+2$ bins (7:383-385, 28:57-59). See lines 1470 through 1960.

After the voltage samples were sorted into bins, a chi-square test, which included Yate's correction, was performed at the 0.05 level of significance (13:562-567). In the event the number of degrees of freedom was greater than 30, the 0.05 probability value of χ^2 was found by

$$\chi^2_{0.05} = \frac{\left(\sqrt{2v-1}+1.645\right)^2}{2} \tag{6-3}$$

where v is the number of degrees of freedom (44:106-107). The expected frequencies in the chi-square test were computed using an error function algorithm devised by Beaulieu (5)

$$Q(x) = \frac{1}{2} - \frac{2}{\pi} \cdot \left[\sum_{\substack{n=1 \\ n \text{ odd}}}^{33} \frac{\exp\left[-\left(\frac{n\pi}{14\cdot\sqrt{2}}\right)^2\right] \cdot \sin\left(\frac{n\pi}{14}\cdot x\right)}{n} \right] \tag{6-4}$$

See lines 3390 through 3810 and lines 4440 through 4520 of the Turner noise quality program.

The theory on FM-by-noise predicts that the grouped voltage samples should pass the chi-square test as the DVR is held constant and the NVR or, equivalently, the SVR is increased. Specifically, receiver noise resulting from FM-UBN and FM-WBN is theoretically Gaussian and should pass the chi-square test. Such was the case. See Tables C-1 and C-2 in Appendix C.

Finally, Turner noise quality was calculated. Initial trial runs and preliminary results indicate that the ability to measure Turner noise quality has been re-established. The results are presented in Tables C-1 and C-2. Note that for a fixed receiver bandwidth, the Turner noise quality tends to increases as the SVR increases.

6.4 Two Alternative Noise Quality Measures Proposed

Two alternative noise quality measures are proposed in this section: *IF noise quality* and *RF noise quality*. The IF noise quality coefficient or factor can be used as a stand-alone description of a noise jammer's effectiveness, or it can be used to modify existing formulae that describe jamming effectiveness. Like Turner noise quality, IF noise quality uses an error measure to assess a time-domain penalty for non-normality of the victim receiver output. Unlike Turner noise quality, IF noise quality assesses a frequency-domain penalty on jammer effectiveness for non-flatness of the victim receiver output in the frequency-domain. Turner *et al.* did not measure PSD flatness because, as they explain, "...[w]ith very few exceptions, the noise produced in realistic receiver bandwidths by operational or developmental noise jammers has been observed to be almost white during the five-year period of sponsored investigations" (43:118).

The second noise quality factor, RF noise quality, measures the baseband normality and frequency-domain impulsivity of WBFM-by-noise barrages, modulated by Gaussian noise, at RF. Recall the baseband normality, in terms of Woodward's theorem, and frequency-domain impulsivity of FM-by-noise barrages was discussed in Chapter V and illustrated in Figures 5-3 through 5-5. This noise quality measure is proposed because it is hypothesized that the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF can be correlated with jamming effectiveness at IF. That is, the effectiveness of a given barrage, in terms of noise jamming, should increase as the normality of the modulating or baseband noise process increases and the frequency-domain impulsivity of the barrage decreases. If this hypothesis is true, then noise jammer effectiveness can be evaluated at RF with frequency-domain measurements alone rather than at IF with both time- and frequency-domain measurements.

6.4.1 Noise Quality at IF. This is the first of two alternative noise quality measures proposed in this thesis. This first alternative noise quality measure is an attempt at improving upon the concepts developed by Turner *et al.*, and will be referred to as IF noise quality. A complete listing of the program used to compute IF noise quality is contained in Appendix A.

IF noise quality, as defined in this investigation, assesses two penalties on the resultant jamming at the output of the front end of a receiver being subjected to noise jamming. The first penalty, p_t, is assessed for non-normality in the time-domain, and the second penalty, p_f, is assessed for non-flatness in the frequency-domain. The two penalties are then used to calculate IF noise quality, ρ_{IF}. The methods of assessing the time- and frequency-domain penalties, specific to this investigation, are described next, along with a detailed explanation of the IF noise quality program listing contained in Appendix A.

In terms of data-acquisition and data-processing, the Turner noise quality program and the IF noise quality program are identical up to the point where the chi-square goodness-of-fit test is conducted. Hence, the discussion of data-acquisition and data-processing issues presented in the section describing the Turner noise quality program also applies to IF noise quality.

The time-domain penalty for non-normality is calculated in lines 1820 through 1950 of the IF noise quality program. The algorithm is loosely based on one presented by Shanmugan and Briephol

(35:497–500). The program transforms the array containing the histogram data into a sequence of sample probability density estimates defined at the midpoints of the inner K equal-width bins (7:383–385). An error measure is obtained by subtracting the sample probability density estimate associated with each bin from the corresponding theoretical value of the ideal normal density, having the same mean and variance, evaluate at the same midpoint. The result is squared and divided, or normalized, by the theoretical value squared. The normalized error associated with each of the K equal-width bins is summed, and an average error is calculated by dividing the sum by K. This average error is the time-domain penalty p_t. Mathematically, p_t can be written

$$p_t = \frac{1}{K}\sum_{i=1}^{K}\sqrt{\frac{\left(\left(K \cdot N_i\right)/N \cdot (B-A) - p_G(x_{ci})\right)^2}{p_G^2(x_{ci})}} \tag{6-5}$$

where

A = lower edge of first equal width bin
B = upper edge of K - th equal width bin
K = number of equal width bins
N = number of voltage samples
N_i = number of voltage samples in i -th bin
x_{ci} = midpoint of i - th bin
$p_G(x_{ci})$ = value of Gaussian pdf at midpoint of i - th bin

Note the quantity $(B-A)/K$ is simply the bin width. This definition of a time-domain penalty seemed justified in light of the references cited and in light of the various error measures and vector and matrix norms described in the literature (7:22-24; 45:87–94). The algorithm produced results consistent with the theory on FM-by-noise, assessing increasingly lenient penalties for jamming that approached or surpassed the requirements of WBFM-WBN. See Tables C-3 and C-4.

The frequency-domain penalty p_f is defined as the ratio of measured jamming power to ideal jamming power referred to the 3 dB bandwidth of the receiver:

$$p_f = \frac{\text{measured power}}{\text{ideal power}} \tag{6-6}$$

In this investigation, an HP 8566B programmable spectrum analyzer was used to make power measurements. The trace values, in dBm/Hz, were read into an array, converted to milliwatts/Hz, and then added to approximate the power contained in the jamming. This algorithm proved to be quite similar to one used by the designers of the spectrum analyzer (42). However their algorithm took the noise bandwidth of the spectrum analyzer into consideration. This consideration was not necessary in the program because the frequency-domain penalty is a power ratio. The noise bandwidth would have appeared in the numerator and denominator of the ratio and canceled out.

The ideal jamming power was defined in terms of the maximum PSD magnitude of a given trace. That is, the ideal jamming power was calculated by finding the power contained in a theoretical trace that had a uniform magnitude, equal to the maximum magnitude of the measured trace, across the 3 dB bandwidth of the victim receiver. In short, the ideal jamming power was calculated by assuming that the noise was truly flat or white over the 3 dB bandwidth of the victim receiver. See lines 3260 to lines 3380.

Note that this penalty is conservative because it is based on the erroneous assumption that the ideal jamming power can be uniform across the 3 dB bandwidth of the victim receiver. The assumption is erroneous because the 3 dB bandwidth of a filter is, by definition, non-uniform. For absolute precision and accuracy, the ideal jammer power should have been defined in terms of the contour of the 3 dB bandwidth of the filter being jammed instead of a theoretical, but unrealizable, uniform contour. Unfortunately, time constraints prevented this idea from being fully explored. As a result of the erroneous assumption on which the frequency-domain penalty is based, p_f will always be less than one. Hence, the penalty is conservative.

The algorithm for assessing the frequency-domain penalty, p_f, produced inconsistent results. See Tables C-3 and C-4. The frequency-domain penalty, for a fixed DVR, should have decreased, with increasing SVR. That is, the value of p_f should have *increased* and approached unity with increasing SVR. However, p_f did not consistently increase with increasing SVR. This inconsistency was probably due to the trace averaging provided by the video bandwidth selected on the HP 8566B spectrum analyzer and the amount of post-sweep smoothing provided by the program. See Appendix C for spectrum analyzer settings and line 430 of the IF noise quality program for the argument used with the HP 8566B SMOOTH function.

Finally, the program computes the IF noise quality, ρ_{IF}:

$$\rho_{IF} = (1 - p_i) \cdot p_f \tag{6-7}$$

Results from preliminary tests indicate that the IF noise quality measure proposed in this investigation may be useful in measuring the effectiveness of FM-UBN and FM-WBN jamming scenarios given that the DVR is equal to or greater than one. Since FM-LFN is a deception-type jamming, the utility of measuring the IF noise quality of FM-LFN scenarios is questionable. See Tables C-3 and C-4.

The IF noise quality coefficient or factor, $\rho_{IF,}$ can be used as a stand-alone description of a noise jammer's effectiveness, or it can be used to modify existing formulae that describe jamming effectiveness. For example, the ultimate effect of noise jamming is to increase the noise figure of a victim receiver by increasing its equivalent device temperature (16:158). Jammer temperature, T_J, as defined in Barton's text, measures the increase in equivalent device temperature or effective input temperature produced by noise jamming (4:139). This quantity, T_J, could be multiplicatively scaled by ρ_{IF}:

$$T_J = \frac{\rho_{IF} P_j G_j A_r F_j^2}{4\pi k_B B_J R_j^2 L_{aj}} \tag{6-8}$$

where

ρ_{IF} = IF noise quality factor

$P_j G_j$ = effective radiated power of the jammer

A_r = effective aperature area of radar antenna

F_j = the pattern - propagation factor of the radar receiving antenna in the jammer direction

k_B = Boltzmann's constant

B_j = bandwidth of jammer spectrum

R_j = jammer range

L_{aj} = one - way propagation loss from jammer

Additionally, link budget calculations could be appropriately modified by the IF noise quality factor (16:163-169). As a final suggestion, ρ_{IF} could be used to weight Shannon's definition of equivocation or Shannon's channel-capacity formula.

6.4.2 Noise Quality at RF. This section proposes a noise quality measure, referred to as RF noise quality, that measures the baseband normality and frequency-domain impulsivity of WBFM-by-noise barrages, modulated by Gaussian noise, at RF. This noise quality measure is proposed because it is hypothesized that the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF can be correlated with jamming effectiveness at IF. This section describes RF noise quality in detail. A complete program listing of RF noise quality as implemented for this investigation is contained in Appendix A. RF noise quality was measured with an HP 8566B programmable spectrum analyzer.

The program first prompts the user for the peak frequency deviation of the barrage. An UNKNOWN option is provided because the peak frequency deviation is usually not known when testing operational jammers, only the barrage width is known. If the UNKNOWN option is selected, the user is prompted for the barrage width. Further instructions follow these first prompts. Next, the program commands the HP 8566B programmable spectrum analyzer to display either the FM bandwidth, defined by Eq. 3-10, or the barrage bandwidth input by the user.

As an aside, even if the peak frequency deviation of a barrage or FM signal is unknown, it can be measured with a spectrum analyzer. Simply increase the resolution bandwidth until the trace is very broad or thick. Next, tune the center frequency of the spectrum analyzer so that thickest portion of the trace is centered on the display. The thickness or width of the centered portion of the trace is approximately twice the peak frequency deviation (19:29).

Other methods of specifying the barrage bandwidth were considered. For example, the barrage bandwidth was defined 3 dBm up from the noise floor. Marker functions were used to zoom in on and display the barrage width thus defined. However, remote operation of the HP 8566B spectrum analyzer markers was not straightforward. Hence, this approach was abandoned. The user may wish to try some other definition of barrage width and automatically zoom in on and display it. As an alternative approach, the user may want to try a different marker scheme or the PWRBW function available on the HP 8566B spectrum analyzer. The PWRBW function returns the bandwidth which contains the user-specified percentage of the total power under the trace. Thus, the user can define the barrage bandwidth as that

bandwidth of the signal which contains, say, 90% of the power in the signal. Ultimately, the FM bandwidth as defined by Carson's Rule, Eq. 3-10, was chosen because it is the most common definition of FM bandwidth.

After the FM bandwidth is displayed on the spectrum analyzer, the program computes the time-domain penalty, p_t. It is important to note that definition of p_t was based on the assumption that the RF carrier was frequency-modulated by white Gaussian noise. This assumption was made because it has been shown that white, Gaussian noise is the most unpredictable of all noise processes (36:21). Furthermore, it has also been shown that erfed-noise barrages and uniform PSD jamming barrages do not necessarily produce a greater jamming effect at IF than non-erfed-noise barrages and non-uniform PSD jamming barrages (8:14-30; 46). Because of these facts, it is reasonable to assume that, given a specific WBFM-WBN jamming scenario, an FM-by-noise barrage modulated by bandlimited, white Gaussian noise will produce effective jamming in its intended victim receiver. Hence, the algorithm for the time-domain penalty was written based on the assumption that the RF carrier was frequency-modulated by white Gaussian noise.

In the case of WBFM-by-noise, modulated by Gaussian noise, Woodward's Theorem predicts that the PSD of the barrage, or simply the barrage, at RF should have the shape of the probability density associated with the underlying baseband noise. Assuming the baseband noise process is Gaussian, the RF barrage should be bell-shaped or normal. Hence, the shape of the RF barrage can be quantitatively compared to a bell-shaped or normal curve. This quantitative comparison yields a value similar to the time-domain penalty assessed in measuring IF noise quality. For the sake of continuity, this quantity is also referred to as a time-domain penalty. In short, the time-domain penalty associated with RF noise quality measures the normality of the underlying baseband noise. Note that Woodward's theorem predicts nothing about the PSD of the underlying baseband noise, and the time-domain penalty does not measure the uniformity of the baseband noise PSD.

The HP 8566B programmable spectrum analyzer used to measure RF noise quality provided another method of gauging the normality of the baseband noise. The HP 8566B spectrum analyzer had a function called PDF that measured the probability density of the instantaneous frequency. Specifically, the HP

8566B could be programmed to produce a histogram of the instantaneous frequencies, or frequency components of the barrage, that had magnitudes above a user-defined threshold. This second method of gauging the normality of the baseband noise could not be fully explored because of time constraints.

The reader may be wondering why the normality of the baseband noise and the baseband noise PSD were not measured directly. The reason is that in most operational jammers it is impractical or impossible to make direct time-and frequency-domain measurements on the baseband noise. Hence, some indirect method of measuring the normality of the baseband had to be devised.

Before the shape of the actual barrage can be compared to the ideal bell-shaped barrage, the actual barrage must be mathematically characterized. The algorithm that characterizes the barrage follows. Dimensions in the equations that follow are bracketed. It can be assumed that the RF barrage has been mixed down to baseband, regardless of the actual RF center frequency. Hence, the actual barrage can be thought of as a double-sided PSD centered at 0 Hz.

Since it is assumed that the barrage is a double-sided PSD centered about 0 Hz and since it is also assumed that the barrage is modulated by Gaussian noise, the actual or measured barrage may be written

$$\hat{S}(n,\sigma)\left[\frac{\mathrm{V}}{\mathrm{Hz}}\right]=\frac{\alpha}{\sqrt{2\pi(\sigma[\mathrm{Hz}])^2}}\cdot\exp\left[-\frac{1}{2}\cdot\left(\frac{n[\mathrm{Hz}]}{\sigma[\mathrm{Hz}]}\right)^2\right][\mathrm{V}] \tag{6-9}$$

where

σ = parameter to be estimated

n = the value of the display ordinate axis in Hertz [Hz]

$\hat{S}(n,\sigma)$ = value of measured PSD in volts / Hz [V / Hz]

and α is a scaling factor. See, for example, the Manual Display in Fig. 5-5. Note that $\hat{S}$ is a function of both n and σ; the latter parameter must be estimated from the actual barrage. The choice of the discrete variable n for the value of the display ordinate axis will be explained momentarily. The program downloads the barrage values in V/Hz, and the scaling factor α is found by setting n to zero. Thus,

$$\alpha=\sigma\cdot S_0\cdot(2\pi)^{\frac{1}{2}} \tag{6-10}$$

where $\hat{S}_o$ is the value of Eq. 6-9 when $n=0$ Hz. Eq. 6-10 can be used to rewrite Eq. 6-9 as

$$\hat{S}(n,\sigma)\left[\frac{\mathrm{V}}{\mathrm{Hz}}\right]=\hat{S}_o\cdot\exp\left[-\frac{1}{2}\cdot\left(\frac{n[\mathrm{Hz}]}{\sigma[\mathrm{Hz}]}\right)^2\right]\left[\frac{\mathrm{V}}{\mathrm{Hz}}\right] \tag{6-11}$$

Eq. 6-11 is used to estimate σ. Some algebraic manipulation yields

$$\frac{\hat{S}(n,\sigma)}{\hat{S}_o}=\exp\left[-\frac{1}{2}\cdot\left(\frac{n[\mathrm{Hz}]}{\sigma[\mathrm{Hz}]}\right)^2\right] \tag{6-12}$$

Taking the natural logarithm of both sides of Eq. 6-12 and performing a bit more algebra yields

$$\sigma^2=-\frac{n^2}{2}\Big/\ln\left(\frac{\hat{S}(n,\sigma)}{\hat{S}_o}\right) \tag{6-13}$$

Thus, a maximum-likelihood estimate, s^2, of σ^2 can be computed from

$$s^2=\frac{1}{1000}\sum_{n=-500}^{500}\frac{n^2}{2}\Big/\ln\left(\frac{\hat{S}_o}{\hat{S}_n}\right),\quad n\neq 0 \tag{6-14}$$

where $\hat{S}_n$ is the n-th value of the measured barrage. This is simply the mean of σ^2. Note that since each HP 8566B trace contains 1001 *discrete* data points and since it is assumed that barrage is a double-sided PSD centered at 0 Hz, n is a discrete variable in Eq. 6-9, and n ranges from -500 Hz to 500 Hz in Eq. 6-14. However, since Eq. 6-14 is undefined at $n=0$, this data point is omitted; recall $n=0$ was used to find the scaling factor, α. The number of data points used to estimate s^2 can be modified by changing the value of the variable named *Limit* in the program, the step size of the FOR loop, or both. Note also that it is *not* necessary to scale n so that it corresponds to the actual bandwidth displayed on the spectrum analyzer as long as n is not scaled when computing the values associated with the ideal bell-shaped barrage.

The ideal bell-shaped barrage, $S_I(n)$ is computed using s^2:

$$S_I(n)=\hat{S}_o\cdot\exp\left[-\frac{1}{2}\cdot\left(\frac{n}{2}\right)^2\right],\qquad -500\leq n\leq 500 \tag{6-15}$$

The shape of the actual barrage is compared to this ideal barrage, and, as was done in IF noise quality, a

time-domain penalty is computed:

$$p_t = \frac{1}{1001}\sum_{n=-500}^{500}\sqrt{\frac{\left(\hat{S}_n - S_I(n)\right)^2}{S_I^2(n)}} \tag{6-16}$$

Next, a frequency-domain penalty, p_f, can be assessed in a manner similar to that used in measuring IF noise quality. See the Eq. 6-6 and the discussion about the equation. Specifically, the power contained in the 3 dB bandwidth of the barrage is divided by the ideal power contained in the same bandwidth had the barrage been truly flat or white across that 3 dB bandwidth. Note that this penalty is conservative in that a 3 dB bandwidth, by definition, can never be flat. Hence, the frequency-domain penalty as defined and assessed in this noise quality measurement, like the penalty assessed in IF noise quality, will always be less than unity. The ideal power is defined, lines 920 through 960, in terms of the maximum amplitude measured in the full bandwidth of the barrage. The maximum amplitude of the barrage is measured and then later multiplied by 1001 because that is the number of points in a trace.

Note that the resolution bandwidth is queried from the spectrum analyzer and saved immediately after the maximum amplitude is acquired. The resolution bandwidth changes with spanwidth, and as the resolution bandwidth decreases, so to does the overall magnitude of the trace. When the program commands the spectrum analyzer to tune the 3 dB bandwidth in lines 970 through 1150, the resolution bandwidth decreases. Therefore, to make a fair power comparison, the baseline resolution bandwidth has to be retuned, line 1300. If this precaution had not been taken, the frequency-domain penalty would always have been artificially high.

It is important to note that frequency-domain penalty as defined in this section is different from the frequency-domain penalty defined in the section on IF noise quality. The frequency-domain penalty assessed in measuring RF noise quality gives a quantitative indication of how impulsive the FM-by-noise barrage is at RF in the frequency-domain. It does not give an indication of how flat or white the baseband noise process is. Finally, the RF noise quality ρ_{RF} is defined, like ρ_{IF}:

$$\rho_{RF} = (1 - p_t)\cdot p_f \tag{6-7}$$

Both Turner noise quality and IF noise quality measure the effectiveness of a given noise jammer at the front end of a simulated victim receiver. Specifically, Turner noise quality and IF noise quality are measured at the output of the IF filter of the victim receiver. Measuring these two noise quality factors is time-consuming because Turner noise quality and IF noise quality are based, in large part, on the non-trivial statistical analysis of long records of time-series data. Just acquiring the data is time-consuming.

RF noise quality, on the other hand, measures the baseband normality and the frequency-domain impulsivity of WBFM-by-noise barrages, modulated by Gaussian noise, at RF. Recall the baseband normality and frequency-domain impulsivity of FM-by-noise barrages were discussed in Chapter 5 and illustrated in Figures 5-3 through 5-5. That is, the effectiveness of a given barrage, in terms of noise jamming, should increase as the normality of the baseband noise process increases and the frequency-domain impulsivity of the barrage decreases. If this hypothesis is true, then it may be possible to evaluate noise jammer effectiveness at RF with frequency-domain measurements alone rather than at IF with both time- and frequency-domain measurements. More specifically, it may be possible to correlate RF noise quality with Turner noise quality and IF noise quality and, thus, measure the effectiveness of different jammers, against the same victim receiver of bandwidth B_V, without simulating the victim receiver. Furthermore, since RF noise quality relies on frequency-domain measurements alone, it can be measured more quickly and simply than both Turner noise quality and IF noise quality.

Results from preliminary tests might lead one to believe that the time-domain penalty associated with RF noise quality factor increases as the deviation ratio D decreases. See Tables C-5 through C-7. However, baseband normality is not proportional to baseband noise bandwidth, and the fact that the time-domain penalty, p_t, became more lenient with decreasing D in this investigation does not imply that the baseband noise became more Gaussian with decreasing D. Recall, FM-by-noise jamming becomes less frequency-domain impulsive as the deviation ratio becomes smaller. The time-domain penalty decreased with decreasing deviation ratio because the spectrum analyzer traces on which the time-domain penalties were based became more well-defined with decreasing deviation ratio. See the Manual Displays shown in Figures 5-3 through 5-5. A meaningful comparison of time-domain penalties can be made when barrages with

equal deviation ratios are generated with different noise generators. Overall, the RF noise quality measure produced results consistent with the theory.

6.5 Summary

Noise quality, as a measure of noise jammer effectiveness, was revisited in this chapter. Literature directly and indirectly related to noise quality was reviewed. The noise quality work done by Turner *et al.* was discussed at length, and, at the request of the sponsors of this investigation, the experimental work of Turner *et al.* was reproduced. Preliminary results indicate that the ability to measure Turner noise quality has been re-established.

In addition to the review and reproduction of the work done by Turner *et al.*, two original noise quality measures were proposed. The first was called *IF noise quality*. Like Turner noise quality, IF noise quality used an error measure to assess a time-domain penalty against the output of a victim receiver. This penalty was assessed for non-normality in the time-domain. Unlike Turner noise quality, IF noise quality assessed a frequency-domain penalty. This penalty was assessed against the output of a victim receiver for non-uniformity of its PSD. The second noise quality factor, *RF noise quality*, measured the baseband normality and frequency-domain impulsivity of WBFM-by-noise barrages, modulated by Gaussian noise, at RF. This noise quality measure was proposed because it is hypothesized that the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF can be correlated with jamming effectiveness at IF.

VII. Conclusions and Recommendations

7.1 Conclusions

7.1.1 FM-by-Noise. From an analytical perspective, this investigation presented a unique description of FM-by-noise which included a thorough review, consolidation, and elucidation of the existing theory on the topic. The description of FM-by-noise presented in this thesis developed three new ratios: the deviation-to-victim ratio (DVR), the noise-to-victim ratio (NVR), and the sweep-to-victim ratio (SVR). These ratios were used to explain and predict the behavior of FM-by-noise at both RF and IF. From an experimental perspective, this investigation presented results obtained from FM-by-noise jamming experiments conducted using commercial test and measurement equipment. Specifically, the time- and frequency-domain behavior of FM-by-noise jamming at RF and IF, in terms of the DVR, NVR, and SVR, was studied with the baseband noise bandwidth, peak frequency deviation, and victim receiver bandwidth as parameters. An important feature of the experimental portion of this investigation was the inclusion of computer-generated reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise at RF and IF. The behavior of FM-by-noise illustrated in the oscilloscope and spectrum analyzer displays was related back to the description of FM-by-noise jamming developed in the analytical portion of the investigation.

It was shown that any meaningful discussion of FM-by-noise must include specific references to at least three ratios: the deviation ratio or *D*, the peak frequency deviation-to-victim receiver bandwidth ratio or DVR, and the noise bandwidth-to-victim receiver bandwidth ratio or NVR. A fourth ratio, the sweep-to-victim receiver bandwidth ratio or SVR, defined as the product of the NVR and DVR, was found to be useful in discussing FM-by-noise jamming scenarios.

The deviation ratio, *D*, quantitatively indicates the degree to which the frequency-modulation process spreads the baseband signal at RF. FM signals are classified as NBFM, UDRFM, or WBFM, depending on the numerical value of the deviation ratio associated with a given FM signal. FM-by-noise barrages are usually WBFM signals. In terms of FM-by-noise, the deviation ratio gives a quantitative indication of

the frequency-domain impulsivity of a given FM-by-noise barrage. The experiments illustrated and explained the frequency-domain impulsivity of FM-by-noise jamming.

The second ratio, the DVR, tells how well a given FM-by-noise barrage fills a given victim receiver. If the DVR is less than one, the barrage will be too small to produce noise jamming, and the receiver will see a noisy sinusoid. On the other hand, if the DVR is too large, the frequency-modulation process will spread the baseband noise process over too wide a bandwidth, and the jamming deposited in the victim receiver will be ineffective because of its low power level. Ideally, the DVR should be approximately 1.27.

The NVR can be used to predict the behavior of FM-by-noise at IF, *given* that the barrage is WBFM and the DVR is such that the barrage fills the passband of the victim receiver. A particular FM-by-noise jamming scenario can be classified, depending on the NVR of the scenario, into one of three categories: FM-LFN, FM-UBN, or FM-WBN. Th experimental portion of the investigation studied these categories of FM-by-noise jamming. The first category, FM-LFN was shown to have a pulse-like nature at IF. This pulse-like nature of FM-LFN, it was argued, produced a confusion effect by causing numerous false targets of various intensities to appear on the visual display of the victim radar. In contrast to FM-LFN, FM-WBN was shown to repeatedly excite the IF filter of its victim receiver. This repeated excitation produced, in the IF filter of the victim filter, an effect similar to DINA jamming. The experiments showed that FM-UBN behaved much like FM-WBN.

Finally, the SVR gives an indication of how frequently a given jammer will sweep through the passband of its victim receiver and how long these sweeps will last. In the experiments, the SVR was related to the slope of the zero-crossings associated with the baseband noise used to generate the FM-by-noise barrages in the various jamming scenarios investigated.

7.1.2 Noise Quality. The concept of noise quality, a measure of noise jamming effectiveness, was revisited. Three noise quality factors were investigated. The work of Turner *et al.* (43) was reproduced, and two alternative noise quality factors were proposed.

Researchers at Stanford hypothesized that the effectiveness of a given noise jammer could be determined by measuring the first order probability density of the output of a receiver subjected to jamming

from the noise jammer under test and, subsequently, comparing this measured density to an ideal normal density having the same mean and variance. To test their hypothesis, they jammed a receiver and measured the normality of the time-series at the output of the front end of the receiver. They did not directly measure the uniformity of the PSD at the output of the filter. The Stanford team investigated and reported on four tests to measure jammer effectiveness (29).

Turner *et al.* extended the Stanford team's work and demonstrated that the effectiveness of a given noise jammer could be described by a single, albeit *ad hoc*, quantity which they called *noise quality* (43). The noise quality work done by Turner *et al.* was discussed at length, and, at the request of the sponsors of this investigation, the experimental work of Turner *et al.* was reproduced. Preliminary results indicate that the ability to measure Turner noise quality has been re-established.

In addition to the review and reproduction of the work done by Turner *et al.*, two original noise quality measures were proposed. The first was called *IF noise quality*. Like Turner noise quality, IF noise quality used an error measure to assess a time-domain penalty. This penalty was assessed against the output process of a victim receiver for non-normality in the time-domain. Unlike Turner noise quality, IF noise quality assessed a frequency-domain penalty against the output process of a victim receiver. This penalty was assessed for non-uniformity of the output PSD.

In preliminary tests, the algorithm used to compute the time-domain penalty for IF noise quality produced consistent results; however, the algorithm used to compute the frequency-domain penalty produced inconsistent results. See Tables C-3 and C-4. Despite this inconsistency, the IF noise quality measure produced meaningful results for FM-UBN and FM-WBN jamming scenarios. Finally, results from preliminary tests indicate that the IF noise quality measure, as defined in this investigation, should not be to measure FM-LFN jamming scenarios.

The second noise quality factor, *RF noise quality*, measured the baseband normality and frequency-domain impulsivity of WBFM-by-noise barrages, modulated by Gaussian noise, at RF. This noise quality measure was proposed because it was hypothesized that the baseband normality and frequency-domain impulsivity of an FM-by-noise barrage at RF can be correlated with jamming effectiveness at IF. That

is, the effectiveness of a given barrage, in terms of noise jamming, should increase as the normality of the baseband noise process increases and the frequency-domain impulsivity of the barrage decreases. If this hypothesis is true, then it may be possible to evaluate noise jammer effectiveness at RF with frequency-domain measurements alone rather than at IF with both time- and frequency-domain measurements. More specifically, it may be possible to correlate RF noise quality with Turner noise quality and IF noise quality and, thus, measure the effectiveness of different jammers, against the same victim receiver of bandwidth B_V, without simulating the victim receiver. Furthermore, since RF noise quality relies on frequency-domain measurements alone, it can be measured more quickly and simply than both Turner noise quality and IF noise quality. Results from preliminary test are promising. See Tables C-5 through C-7.

7.2 Recommendations

7.2.1 FM-by-Noise–DINA Proof. A rigorous proof of the claim that FM-WBN produces jamming similar to DINA would be a welcome addition to the body of open-literature on FM-by-noise. Doing so, however, requires intimate knowledge of the baseband random process used to generate the FM-by-noise, such as the number of zero-crossings the baseband random process makes, the extent of the deviation above and below zero associated with each zero-crossing, and, finally, the slope of each linearized zero-crossing. This proof should also include precise requirements, in terms of D, the DVR, and the NVR, for FM-by-noise to approximate DINA.

7.2.2 Record Length. In order to reduce processing time, one of the most important questions to answer is what is the shortest record length required to measure normality in the time-domain. A thorough investigation of this question would consider the statistical errors associated with this minimum record length. It would also address the statistical errors associated with this minimum record length and their effect on the measurement of noise quality.

7.2.3 Number of Bins and Bin Width. A related statistical issue is the optimum number of bins needed for a given number of samples or record length. The data collected in this thesis was grouped into K equal-width bins. The width was approximately 0.2 time measured standard deviation. However, it is

generally accepted that equiprobable, rather than equal-width, bins produce better results for the chi-square test (15:69). Perhaps equiprobable bins could be incorporated into the programs.

7.2.4 Spectrum Analyzer Measurements. The algorithm for the frequency-domain penalty in the IF noise quality program produced inconsistent results. These results were calculated on data acquired from the HP 8566B spectrum analyzer. However, all spectrum analyzer measurements were highly dependent on the video bandwidth setting and the argument passed to the HP 8566B SMOOTH function. In this investigation, the video bandwidth setting and the SMOOTH argument were manually selected. Perhaps an algorithm that automatically adjusts the video bandwidth setting and SMOOTH argument to some optimum values for the given measurement can incorporated into the programs.

7.2.5 Noise Quality Validation. The noise quality measurements developed for this thesis need to be validated. A reasonable follow-on thesis, emphasizing practicality, would take, or simulate, several well-characterized FM-by-noise jammers and test their effectiveness using the three noise quality measures discussed in this thesis: Turner noise quality, IF noise quality, and RF noise quality. The thesis would compare the performance of Turner noise quality and IF noise quality as tools to measure jammer effectiveness. Follow-on work would also determine if there is a meaningful and useful correlation between noise quality measured at IF, using both Turner noise quality and IF noise quality, and noise quality measured at RF.

Appendix A Programs

This appendix contains the programs that were written for the experimental portion of the investigation. Each program contains its own documentation. Since Chapter VI discusses noise quality in detail, the reader is urged to consult that chapter before using the noise quality programs presented in this appendix. The code was written in HP Basic specifically for the HP 54111D digitizing oscilloscope and the HP 8566B spectrum analyzer.

A.1 TIMEDMN.BAS

```
10   ! HP BASIC 5.14
20   !
30   ! PROGRAM TIMEDMN.BAS
40   !
50   ! THIS PROGRAM DIGITIZES CHANNEL 1 OF AN HP 54111D OSCILLOSCOPE (ADDRESS 707)
60   ! AND THEN DOWNLOADS THE TRACE TO A DATA FILE  NOTE THAT THE DATA ARE DOWNLOADED
70   ! IN HP ASCII FORMAT. THE DATA MUST BE TRANSLATED TO DOS FORMAT IF IT IS TO BE
80   ! PORTED TO A DOS MACHINE. THE DATA CAN BE TRANSLATED BY USING THE TRANSLATION
90   ! PROGRAM "ASCI2DOS" THAT COMES WITH HP BASIC. JUST "LOAD 'ASCI2DOS'", RUN,
100  ! AND FOLLOW THE INSTRUCTIONS.
110  !
120  OPTION BASE 0
130  MASS STORAGE IS "\BLP\JAM3DATA:DOS,C"              ! CHANGE MASS STORAGE
140  DIM A(500)                                         ! SO FILES ARE DOWNLOADED
150  OUTPUT 707; "ACQUIRE RESOLUTION?"                  ! SET-UP SCOPE TO
160  ENTER 707; Resolution                              ! ACQUIRE TRACES
170  Reso$ = VAL$(Resolution)
180  OUTPUT 707; "ACQUIRE TYPE FILTERED"
190  OUTPUT 707; "ACQUIRE RESOLUTION "&Reso$
200  OUTPUT 707; "ACQUIRE TYPE FILTERED"
210  OUTPUT 707; "ACQUIRE RESOLUTION "&Reso$
220  OUTPUT 707; "WAVEFORM FORMAT ASCII"
230  OUTPUT 707; "DIGITIZE CHANNEL 1"
240  INPUT "What is your filename?",Ascii_file$
250  CLEAR SCREEN
260  ! USE 8 CHARACTER filename WITH 3 CHARACTER ext
270  PURGE Ascii_file$                                  ! CREATE TIME DOMAIN FILE
280  ON ERROR GOTO 290
290  CREATE ASCII Ascii_file$,1
300  ASSIGN @Ascii_file TO Ascii_file$
310  OFF ERROR
320  OUTPUT 707; "WAVEFORM SOURCE MEMORY 5"
330  OUTPUT 707; "YREFERENCE?"
340  ENTER 707; Yref
350  OUTPUT 707; "YINCREMENT?"
360  ENTER 707; Yinc                                    ! GET PARAMETERS FOR USE
370  OUTPUT 707; "YORIGIN?"                             ! DIGITAL TO ANALOG (D/A)
380  ENTER 707; Yorg                                    ! CONVERSION
390  OUTPUT 707; "DATA?"
400  FOR N = 0 TO 500
410   ENTER 707; A(N)                                   ! PROGRAM SOMETIMES
420  NEXT N                                             ! FAILS HERE FOR
```

```
430                                                      ! FOR SOME UNKNOWN REASON
440                                                      ! JUST RE-RUN UNTIL THE
450                                                      ! BUG WORKS ITSELF OUT
460   FOR N = 0 TO 500
470     A(N) = ((A(N)-Yref)*Yinc)+Yorg                   ! PERFORM D/A CONVERSION
480   NEXT N
490   FOR N = 0 TO 500                                   ! OUTPUT TRACE DATA TO FILE
500     OUTPUT @Ascii_file; A(N)
510   NEXT N
420   OUTPUT 707; "LOCAL"                                ! PUT SCOPE IN LOCAL
530   END
```

A.2 FREQDMN..BAS

```
10    !  HP Basic 5.14
20    !
30    ! PROGRAM FREQDMN.BAS
40    !
50    ! THIS PROGRAM DIGITIZES AN HP 8566B SPECTRUM ANALYZER (ADDRESS 718) TRACE
60    ! AND THEN DOWNLOADS THE TRACE TO A DATA FILE. NOTE THAT THE DATA ARE DOWNLOADED
70    ! IN HP ASCII FORMAT. THE DATA MUST BE TRANSLATED TO DOS FORMAT IF IT IS TO BE
80    ! PORTED TO A DOS MACHINE. THE DATA CAN BE TRANSLATED BY USING THE TRANSLATION
90    ! PROGRAM "ASCI2DOS" THAT COMES WITH HP BASIC. JUST "LOAD 'ASCI2DOS'", RUN,
100   ! AND FOLLOW THE INSTRUCTIONS.
110   !
120   MASS STORAGE IS "\BLP:DOS,C"
130   LINPUT "Enter your output filename: filename.ext", Ascii_file$
140   OUTPUT 19; "DEL" Ascii_file$
150   CREATE ASCII Ascii_file$,1
160   ASSIGN @Ascii_file TO "\BLP\"Ascii file$":DOS,C";
170   DIM Trace(1001)
180   OUTPUT 718;"TS;TA;"
190   FOR N=1 TO 1001"
200    ENTER 718; Trace(N)
210    OUTPUT @Ascii_file; Trace(N)
220   NEXT N
230   END
```

A.3 TFDATA.BAS

```
10    ! HP BASIC 5.14
20    !
30    ! PROGRAM TFDATA.BAS
40    !
50    ! THIS PROGRAM SIMULTANEOUSLY DIGITIZES CHANNELS 1 AND 2 OF AN
60    ! HP 54111D OSCILLOSCOPE (ADDRESS 707) AND THEN SWEEPS A TRACE ON
70    ! AN HP 8566B SPECTRUM ANALYZER (ADDRESS 718). THE THREE TRACES, TWO
80    ! FROM THE OSCILLOSCOPE AND ONE FROM THE SPECTRUM ANALYZER, ARE THEN
90    ! DOWNLOADED INTO THREE DATA FILES.  THE USER FIRST IS PROMPTED FOR A
100   ! filename.ext WHERE filename IS 5 CHARACTERS LONG.  SINCE THE
110   ! OSCILLOSCOPE TRACES OR TIME DOMAIN FILES ARE READ FIRST, START THE
120   ! FILENAME WITH, SAY, T. THE NEXT 3 CHARACTERS ARE USER-DEFINED. THE
130   ! 5TH CHARACTER IS CHANGED BY THE PROGRAM TO EITHER A "B" OR AN "O".
140   ! I CHOSE "B" TO SIGNIFY "B"ASEBAND AND CHANNEL ONE, WHILE "O" SIGNIFIES
150   ! THE "O"UTPUT OF A FILTER AND CHANNEL 2. THE PROGRAM CHANGES THE FIRST
160   ! LETTER OF THE FILENAME TO "F" TO DOWNLOAD THE FREQUENCY DOMAIN FILE,
170   ! THAT IS THE SPECTRUM ANALYZER TRACE. THE SIXTH LETTER IN THE filename
180   ! REMAINS "O" FOR CONVENIENCE. THIS FILENAMING ROUTINE IS NOT THE MOST
190   ! ELEGANT, BUT IT WORKS. THE USER IS FREE TO MODIFY THE ROUTINE. NOTE
200   ! THAT THE DATA FILES ARE DOWNLOADED IN HP ASCII FORMAT. THEY MUST BE
```

```
210   ! TRANSLATED TO DOS FORMAT IF THEY ARE TO BE PORTED TO A DOS MACHINE.
220   ! THE DATA CAN BE TRANSLATED BY USING THE TRANSLATION PROGRAM
230   ! "ASCI2DOS" THAT COMES WITH HP BASIC. JUST "LOAD 'ASCI2DOS'", RUN,
240   ! AND FOLLOW THE INSTRUCTIONS.
250   !
260   OPTION BASE 0
270   MASS STORAGE IS "\BLP\JAM3DATA:DOS,C"            ! CHANGE MASS STORAGE
280   DIM A(500)                                       ! SO FILES ARE DOWNLOADED
290   CLEAR SCREEN                                     ! TO A SPECIFIC DIRECTORY
300   PRINT "Enter a FIVE-LETTER filename with THREE-LETTER ext."
310   PRINT ""
320   INPUT "What is your filename?",Ascii_file$
330   CLEAR SCREEN
340   OUTPUT 707; "ACQUIRE RESOLUTION?"                ! SET-UP SCOPE TO
350   ENTER 707; Resolution                            ! ACQUIRE TRACES
360   Reso$ = VAL$(Resolution)
370   OUTPUT 707; "ACQUIRE TYPE FILTERED"
380   OUTPUT 707; "ACQUIRE RESOLUTION "&Reso$
390   OUTPUT 707; "WAVEFORM FORMAT ASCII"
400   REMOTE 718
410   Yes$ = "N"
420   REPEAT                                       ! DIGITIZE O-SCOPE AND ANALYZER
430     OUTPUT 707; "DIGITIZE CHANNEL 1,2"         ! TRACES UNTIL THEY EXHIBIT
440     OUTPUT 718; "S2,TS; O3; TA; "              ! DISTINGUISHING FEATURES
450     INPUT "ARE THE TRACES GOOD?  Y/N",Yes$
460   UNTIL Yes$ = "Y"
470   FOR K = 1 TO 2                               ! DOWNLOAD FILTERED TRACES
480     SELECT K                                   ! FROM MEMORY 5 OR 6
490     CASE = 1                                   ! FOR CHANNEL 1 AND 2
500       Memory$ = "5"                            ! RESPECTIVELY.
510       Ascii_file$[5,5] = "B"                   ! "B" & "O" ARE FOR BOOK-
520     CASE = 2                                   ! KEEPING PURPOSES.
530       Memory$ = "6"                            ! B =  BASEBAND/CHANNEL 1
540       Ascii_file$[5,5] = "O"                   ! O =  FILTER OUTPUT/CHANNEL 2
550     END SELECT
560     ON ERROR GOTO 580
570     PURGE Ascii_file$                          ! CREATE TIME DOMAIN FILE
580     CREATE ASCII Ascii_file$,1
590     ASSIGN @Ascii_file TO Ascii_file$
600     OFF ERROR
610     OUTPUT 707; "WAVEFORM SOURCE MEMORY "&Memory$
620     OUTPUT 707; "YREFERENCE?"
630     ENTER 707; Yref
640     OUTPUT 707; "YINCREMENT?"
650     ENTER 707; Yinc                            ! GET PARAMETERS FOR USE
660     OUTPUT 707; "YORIGIN?"                     ! DIGITAL TO ANALOG (D/A)
670     ENTER 707; Yorg                            ! CONVERSION
680     OUTPUT 707; "DATA?"
690     FOR N = 0 TO 500
700       ENTER 707; A(N)                          ! PROGRAM SOMETIMES
710     NEXT N                                     ! FAILS HERE FOR
720                                                ! FOR SOME UNKNOWN REASON
730                                                ! JUST RE-RUN UNTIL THE
740                                                ! BUG WORKS ITSELF OUT
750     FOR N = 0 TO 500
760       A(N) = ((A(N)-Yref)*Yinc)+Yorg           ! PERFORM D/A CONVERSION
770     NEXT N
780     FOR N = 0 TO 500                           ! OUTPUT TRACE DATA TO FILE
790       OUTPUT @Ascii_file; A(N)
800     NEXT N
810   NEXT K
820   OUTPUT 707; "LOCAL"                          ! PUT SCOPE IN LOCAL
830   DIM Tra(1000)
```

```
840   Ascii_file$[1,1] = "F"                    ! CREATE FREQUENCY DOMAIN FILE
850   ON ERROR GOTO 870
860   PURGE Ascii_file$
870   CREATE ASCII Ascii_file$,1
880   ASSIGN @Ascii_file TO Ascii_file$
890   OFF ERROR
900   FOR N = 0 TO 1000
910     ENTER 718; Tra(N)                       ! DOWNLOAD SPECTRUM ANALYZER
920   NEXT N                                    ! TRACE THAT WAS DIGITIZED
930   FOR N = 0 TO 1000                         ! IN LINE 440
940     OUTPUT @Ascii_file; Tra(N)
950   NEXT N
960   LOCAL 718                                 ! PUT SPECTRUM ANALYZER IN LOCAL
970   MASS STORAGE IS "\BLP:DOS,C"
980   END
```

A.4 Turner Noise Quality

Since this noise quality measure is based on the statistical grouping of voltage samples, the results of this program are dependent upon the number of voltage samples used to calculate Turner noise quality. The total number of samples used depends on the sampling rate, bandwidth of the filter or signal being measured, and the number of memory blocks downloaded to the computer. Note that the total number of samples used is not necessarily equal to the total number of samples acquired from the HP 54111D oscilloscope; correlated samples are ignored. Finally, voltage samples were acquired from the HP 54111D with the RESOLUTION filter set to OFF for quicker data-acquisition times. Additional comments are imbedded in the code. Please read Chapter VI before using this program.

```
10    ! HP BASIC 5.14
20    !
30    ! TURNERNQ.BAS
40    !
50    ! THIS PROGRAM CALCULATES THE NOISE QUALITY AT THE OUTPUT OF A
60    ! FILTER USING THE ALGORITHM DEVELOPED BY TURNER AND OTHERS.
70    ! TURNER, F.M. AND OTHERS. "NOISE QUALITY OPTIMIZES JAMMER
80    ! PERFORMANCE," ELECTRONIC WARFARE/DEFENSE ELECTRONICS, VOL 9:
90    ! 117-122 (NOV/DEC 1977). THIS PROGRAM WAS WRITTEN FOR USE WITH AN
100   ! HP 54111D DIGITIZING OSCILLOSCOPE (ADDRESS 707). AN ADDITIONAL
110   ! FEATURE OF THIS PROGRAM IS THE INCLUSION OF A CHI-SQUARE TEST
120   ! TO TEST FOR NON-NORMALITY.
130   !
140   CLEAR SCREEN
150   KEY LABELS ON
160   OPTION BASE 0
170   BEEP
180   PRINT "SET OSCILLOSCOPE SETTINGS FOR SLOWEST SAMPLING RATE POSSIBLE."
190   PRINT ""
200   PRINT ""
210   INPUT "WHAT IS THE BANDWIDTH OF YOUR SIGNAL/FILTER IN HZ?",Bw
220   CLEAR SCREEN
230   !
240   ! THE VARIABLE Dumps IS USED TO DETERMINE HOW MANY TRACES WILL
```

```
250   ! DOWNLOADED. WHEN DOWNLOADING MULTIPLE TRACES, THE INDIVIDUAL
260   ! TRACES ARE NOT CONTIGUOUS IN TIME BECAUSE OF THE TIME REQUIRED
270   ! TO READ EACH TRACE. THIS TIME DELAY WAS NOT SIGNIFICANT IN THIS
280   ! INVESTIGATION BECAUSE THE TIME DELAY BETWEEN INDIVIDUAL TRACES
290   ! MUCH GREATER THAN THE NOISE CORRELATION DURATIONS ASSOCIATED
300   ! WITH THE SIGNALS/FILTERS USED IN THE EXPERIMENTS. THE USER CAN
310   ! CHANGE THE NUMBER OF TRACES DOWNLOADED BY CHANGING DUMPS.
320   ! DOWNLOADING AND PROCESSING ONE OR TWO TRACES DOES NOT TAKE MUCH
330   ! TIME.
340   !
350   Dumps=2
360   !
370   ! EACH TRACE DOWNLOADED IN NORMAL MODE FROM MEMORY 1 OR 2
380   ! CONTAINS 8192 POINTS. DECLARE ARRAYS TO PROCESS THE DATA.
390   ! Anoise CONTAINS THE SAMPLES. Asorted IS USED LATER IN
400   ! GROUPING THE DATA IN BINS.
410   !
420   Numpoints=Dumps*8192
430   ALLOCATE Anoise(Numpoints-1)
440   ALLOCATE Asorted(Numpoints-1)
450   !
460   ! PERFORM PRELIMINARY OPERATIONS TO DOWNLOAD TRACE DATA.
470   ! HIGHER RESOLUTIONS CAN BE CHOSEN, BUT HIGHER RESOLUTIONS
480   ! DECREASE THE BANDWIDTH SENSITIVITY OF THE OSCILLOSCOPE.
490   ! ALSO TRACE DOWNLOADS ARE FASTER WITH RESOLUTION OFF AND
500   ! CHANNEL DISPLAYS BLANKED.
510   !
520   PRINT "SWEEPING AND ACQUIRING TRACE DATA."
530   PRINT ""
540   OUTPUT 707;"ACQUIRE TYPE NORMAL"
550   OUTPUT 707;"ACQUIRE RESOLUTION OFF"
560   OUTPUT 707;"WAVEFORM SOURCE MEMORY 2"
570   OUTPUT 707;"WAVEFORM FORMAT ASCII"
580   M=0
590   N=0
600   OUTPUT 707;"BLANK CHANNEL 1"
610   OUTPUT 707;"BLANK CHANNEL 2"
620   REPEAT
630     IF Dumps>1 THEN PRINT TAB(10);"ACQUIRING TRACE ";M+1
640     Counter=0
650     OUTPUT 707;"DIGITIZE CHANNEL 2"
660     OUTPUT 707;"DATA?"
670     REPEAT
680       ENTER 707;Anoise(N)
690       Counter=Counter+1
700       N=N+1
710     UNTIL Counter=8191
720     M=M+1
730   UNTIL M=Dumps
740   !
750   ! READ IN PARAMETERS TO PERFORM "DIGITAL TO ANALOG" CONVERSION.
760   ! ALSO READ IN THE TIME DIFFERENCE, Xinc, BETWEEN INDIVIDUAL
770   ! SAMPLES. CALCULATE THE NUMBER OF SAMPLES, Tstep, BETWEEN
780   ! PAIRS OF INDEPEDENT SAMPLES THAT ARE CORRELATED AND CAN BE THROWN
790   ! AWAY. THIS CALCULATION IS BASED ON THE NOISE CORRELATION DURATION
800   ! OF THE SIGNAL, OR EQUIVALENTLY THE FILTER, BEING MEASURED.
810   !
820   OUTPUT 707;"YREFERENCE?"
830   ENTER 707;Yref
840   OUTPUT 707;"YINCREMENT?"
850   ENTER 707;Yinc
860   OUTPUT 707;"YORIGIN?"
870   ENTER 707;Yorg
```

```
880   OUTPUT 707;"XINCREMENT?"
890   ENTER 707;Xinc
900   Tstep=INT(1/(Bw*Xinc))+1
910   !
920   ! REDISPLAY CHANNELS AND PUT OSCOPE IN LOCAL.
930   !
940   OUTPUT 707;"VIEW CHANNEL 1"
950   OUTPUT 707;"VIEW CHANNEL 2"
960   OUTPUT 707;"LOCAL"
970   !
980   ! PERFORM ARRAYS OPERATIONS TO GROUP INDEPENDENT SAMPLES.
990   !
1000  Counter=0
1010  FOR N=0 TO Numpoints-1 STEP Tstep
1020    Anoise(N)=Anoise(N)
1030    Asorted(N)=Anoise(N)
1040    Counter=Counter+1
1050  NEXT N
1060  Numpoints=Counter
1070  CLEAR SCREEN
1080  !
1090  PRINT "PERFORMING DIGITAL TO ANALOG CONVERSION."
1100  FOR N=0 TO Numpoints-1
1110    Anoise(N)=((Anoise(N)-Yref)*Yinc)+Yorg
1120    Asorted(N)=Anoise(N)
1130  NEXT N
1140  !
1150  ! PERFORM ARRAY OPERATIONS TO FORM ARRAYS OF UNCORRELATED DATA.
1160  !
1170  ALLOCATE Noise(Numpoints-1)
1180  ALLOCATE Sorted(Numpoints-1)
1190  FOR N=0 TO Numpoints-1
1200    Noise(N)=Anoise(N)
1210    Sorted(N)=Asorted(N)
1220  NEXT N
1230  DEALLOCATE Anoise(*)
1240  DEALLOCATE Asorted(*)
1250  CLEAR SCREEN
1260  !
1270  ! ESTIMATE MEAN AND VARIANCE AND NORMALIZE TO ZERO-MEAN
1280  ! UNIT VARIANCE.
1290  !
1300  PRINT "ESTIMATING PARAMETERS."
1310  Xbar=0
1320  FOR N=0 TO Numpoints-1
1330    Xbar=Noise(N)+Xbar
1340  NEXT N
1350  Xbar=Xbar/(Numpoints)
1360  S2=0
1370  FOR N=0 TO Numpoints-1
1380    S2=S2+(Noise(N)-Xbar)^2
1390  NEXT N
1400  S2=S2/(Numpoints-1)
1410  S=SQRT(S2)
1420  FOR N=0 TO Numpoints-1
1430    Noise(N)=(Noise(N)-Xbar)/S
1440    Sorted(N)=(Sorted(N)-Xbar)/S
1450  NEXT N
1460  !
1470  ! SORT VOLTAGES INTO BINS. Sorted(*) IS USED TO DETERMINE
1480  ! THE WIDTH OF THE TWO END BINS. NOTE ALL BINS EXCEPT THE
1490  ! THE TWO END BINS ARE OF EQUAL WIDTH. THE PROGRAM FORCES
1500  ! THE TWO END BINS TO CONTAIN AT LEAST 2 TO 3 SAMPLES.
```

```
1510  ! Left AND Right ARE USED TO DEFINE WINDOW BOUNDARIES FOR
1520  ! THE HISTOGRAM DISPLAY.
1530  !
1540  Left=MIN(Noise(*))
1550  Right=MAX(Noise(*))
1560  MAT SORT Sorted(*)                ! DETERMINE START AND END POINTS
1570  Amark=2                           ! FOR K EQUAL WIDTH BINS
1580  A=Sorted(Amark)                   ! AND SORT VOLTAGES INTO BINS
1590  Bmark=Numpoints-2
1600  B=Sorted(Bmark)
1610  IF (A=Left) THEN
1620    Counter=Amark                   ! FORCE LEFT-END AND RIGHT-END
1630    REPEAT                          ! BINS TO CONTAIN AT LEAST TWO
1640      Counter=Counter+1             ! SAMPLES AND BE OF NON-ZERO WIDTH
1650      A=Sorted(Counter)
1660    UNTIL A>Sorted(Counter-1)
1670  END IF
1680  IF (B=Right) THEN
1690    Counter=Bmark
1700    REPEAT
1710      Counter=Counter-1
1720      B=Sorted(Counter)
1730    UNTIL B<Sorted(Counter+1)
1740  END IF
1750        ! CALCULATE BIN WIDTH, Delta, BASED ON INTEGER NUMBER OF
1760        ! EQUAL WIDTH BINS. NOT RECALCULATING DELTA BASED ON
1770        ! AN INTEGER K CAUSES PROBLEMS WITH PLOTTING THE HISTOGRAM.
1780        !
1790  Delta=.2
1800  K=(B-A)/Delta
1810  K=INT(K)
1820  Delta=(B-A)/K
1830  ALLOCATE Pdf(K+1)
1840  CLEAR SCREEN
1850  PRINT "SORTING VOLTAGES INTO BINS."
1860  FOR M=0 TO Numpoints-1
1870    SELECT Noise(M)
1880    CASE <A
1890      Pdf(0)=Pdf(0)+1
1900    CASE >=B
1910      Pdf(K+1)=Pdf(K+1)+1
1920    CASE ELSE
1930      J=INT((Noise(M)-A)/Delta)+1
1940      Pdf(J)=Pdf(J)+1
1950    END SELECT
1960  NEXT M
1970  CLEAR SCREEN
1980  !
1990  ! PRINT STATISTICAL INFORMATION
2000  !
2010  Dxbar=PROUND(Xbar,-4)
2020  Ds=PROUND(S,-4)
2030  Dl=PROUND(Left,-4)
2040  Dr=PROUND(Right,-4)
2050  Mf=MIN(Pdf(*))
2060  Xf=MAX(Pdf(*))
2070  CLEAR SCREEN
2080  PRINT "SAMPLE MEAN =";Dxbar,TAB(40),"SAMPLE STANDARD DEVIATION =";Ds
2090  PRINT ""
2100  PRINT "MINIMUM VOLTAGE = ";Dl,TAB(40),"MAXIMUM VOLTAGE =";Dr
2110  PRINT ""
2120  PRINT "MINIMUM FREQUENCY =";Mf,TAB(40),"MAXIMUM FREQUENCY =";Xf
2130  PRINT ""
```

```
2140 PRINT "NUMBER OF POINTS = ";SUM(Pdf),TAB(40),"NUMBER OF INTERVALS = ";K+2
2150 PRINT ""
2160 PRINT "NOISE CORRELATION DURATION GTE TO ";1/(Bw)
2170 PRINT ""
2180 PRINT "TIME BETWEEN DECORRELATED SAMPLES IS APPROX ";(Tstep)*Xinc
2190 PRINT ""
2200 PRINT "BIN WIDTH = ";PROUND(Delta*S,-4)
2210 PRINT ""
2220 PRINT ""
2230 PRINT ""
2240 PRINT TAB(25),"Press F2 to CONTINUE"
2250 PAUSE
2260 CLEAR SCREEN
2270 !
2280 ! CALCULATE THEORETICAL PDF VALUES AND STORE IN Ideal(*).
2290 !
2300 Low=A
2310 ALLOCATE Ideal(K)
2320 FOR I=1 TO K
2330   Hi=A+I*Delta
2340   X_ci=(Hi+Low)/2
2350   Exp_arg=-.5*(X_ci)^2
2360   Fx_ci=EXP(Exp_arg)/SQRT(2*PI)
2370   Ideal(I)=Fx_ci*Delta*Numpoints
2380   Low=Hi
2390 NEXT I
2400 CLEAR SCREEN
2410 !
2420 ! PLOT THEORETICAL PDF (HISTOGRAM).
2430 !
2440 GINIT
2450 GCLEAR
2460 GRAPHICS ON
2470 KEY LABELS OFF
2480 MOVE 27,90
2490 LABEL "     IDEAL HISTOGRAM"
2500 MOVE 0,70
2510 Label$="Frequency"
2520 FOR I=1 TO 9
2530   LABEL Label$[I,I]
2540 NEXT I
2550 MOVE 55,..
2560 LABEL "Volts"
2570 VIEWPORT 10,120,15,90
2580 FRAME
2590 WINDOW Left,Right,0,MAX(Ideal(*))+5
2600 FOR I=0 TO K+1
2610   SELECT I
2620   CASE =0
2630     MOVE Left,0
2640     RECTANGLE ABS(Left-A),Pdf(0)
2650   CASE =K+1
2660     MOVE B,0
2670     RECTANGLE ABS(Right-B),Pdf(K+1)
2680   CASE ELSE
2690     MOVE A+(I-1)*Delta,0
2700     RECTANGLE Delta,Ideal(I)
2710   END SELECT
2720 NEXT I
2730 PRINT "PRESS F2 TO CONTINUE."
2740 PAUSE
2750 CLEAR SCREEN
2760 KEY LABELS ON
```

```
2770 CLEAR SCREEN
2780 !
2790 ! PLOT HISTOGRAM (PDF ESTIMATE) OF DATA.
2800 !
2810 GINIT
2820 GCLEAR
2830 GRAPHICS ON
2840 KEY LABELS OFF
2850 MOVE 27,90
2860 LABEL "HISTOGRAM OF NOISE VOLTAGES"
2870 MOVE 0,70
2880 Label$="Frequency"
2890 FOR I=1 TO 9
2900   LABEL Label$[I,I]
2910 NEXT I
2920 MOVE 55,10
2930 LABEL "Volts"
2940 VIEWPORT 10,120,15,90
2950 FRAME
2960 WINDOW Left,Right,0,MAX(Pdf(*))+5
2970 FOR I=0 TO K+1
2980   SELECT I
2990   CASE =0
3000     MOVE Left,0
3010     RECTANGLE ABS(Left-A),Pdf(0)
3020   CASE =K+1
3030     MOVE B,0
3040     RECTANGLE ABS(Right-B),Pdf(K+1)
3050   CASE ELSE
3060     MOVE A+(I-1)*Delta,0
3070     RECTANGLE Delta,Pdf(I)
3080   END SELECT
3090 NEXT I
3100 PRINT "PRESS F2 TO CONTINUE."
3110 PAUSE
3120 CLEAR SCREEN
3130 KEY LABELS ON
3140 !
3150 ! CALCULATE EXPECTED FREQUENCIES FOR CHI-SQUARE TEST.
3160 ! TEST IS PERFORMED WITH K+2 BINS. HENCE, K-1 DEGREES
3170 ! OF FREEDOM ARE USED. NOTE USE OF ERF FUNCTION, Fnerf(*).
3180 !
3190 ALLOCATE Expected(K+1)
3200 FOR I=0 TO K+1
3210   Prob=0
3220   SELECT I
3230   CASE =0
3240     X=A
3250     Prob=FNErf(X)
3260     Expected(0)=Numpoints*(1-Prob)
3270   CASE =K+1
3280     X=B
3290     Prob=FNErf(X)
3300     Expected(K+1)=Numpoints*Prob
3310   CASE ELSE
3320     Low_lim=A+((I-1)*Delta)
3330     Up_lim=A+(I*Delta)
3340     Prob=FNErf(Low_lim)-FNErf(Up_lim)
3350     Expected(I)=Numpoints*Prob
3360   END SELECT
3370 NEXT I
3380 !
3390 ! PERFORM CHI-SQUARE TEST. STORE THEORETICAL CHI-SQUARE
```

```
3400  ! VALUES IN Chiray. NOTE THE USE OF YATE'S CORRECTION.
3410  ! ALSO, CHI-SQUARE FOR DEGREES OF FREEDOM GREATER THAN
3420  ! 30 IS CALCULATED BY A FORMULA FOUND IN THE LITERATURE.
3430  !
3440  DIM Chiray(29)
3450  DATA 3.84,5.99,7.81,9.49,11.07,12.59,14.07,15.51,16.92,18.31
3460  FOR N=0 TO 9
3470    READ Chiray(N)
3480  NEXT N
3490  DATA 19.68,21.03,22.36,23.68,25,26.30,27.59,28.87,30.14,31.41
3500  FOR N=10 TO 19
3510    READ Chiray(N)
3520  NEXT N
3530  DATA 32.67,33.92,35.17,36.42,37.65,38.88,40.11,41.34,42.56,43.77
3540  FOR N=20 TO 29
3550    READ Chiray(N)
3560  NEXT N
3570  Chi_square=0
3580  FOR I=0 TO K+1
3590    Term=ABS(Pdf(I)-Expected(I))
3600    Chi_square=Chi_square+((Term-.5)^2)/Expected(I)
3610  NEXT I
3620  Chi_square=PROUND(Chi_square,-2)
3630  IF K-1<=29 THEN
3640    X2=Chiray(K-2)
3650  ELSE
3660    X2=.5*(SQRT(2*(K-1))+1.645)^2
3670  END IF
3680  PRINT "IDEAL CHI-SQUARE = ";PROUND(X2,-2)
3690  PRINT ""
3700  PRINT "CALCULATED CHI-SQUARE = ";Chi_square
3710  PRINT ""
3720  IF (Chi_square<=X2) THEN
3730    PRINT "CHI-SQUARE DOES NOT REJECT THE HYPOTHESIS OF NORMALITY."
3740  ELSE
3750    PRINT "CHI-SQUARE REJECTS THE HYPOTHESIS OF NORMALITY."
3760  END IF
3770  PRINT ""
3780  PRINT ""
3790  PRINT "PRESS F2 TO CONTINUE."
3800  PAUSE
3810  CLEAR SCREEN
3820  !
3830  ! CALCULATE TURNER NOISE QUALITY
3840  !
3850  Mean=0
3860  K1=0
3870  Stdev=0
3880  Var=0
3890  Skewness=0
3900  Kurtosis=0
3910  FOR I=0 TO K+1
3920    Mean=I*Pdf(I)/Numpoints+Mean
3930  NEXT I
3940  FOR I=0 TO K+1
3950    K1=I-Mean
3960    K2=K1^2
3970    Var=K2*Pdf(I)+Var
3980    Skewness=K1*K2*Pdf(I)+Skewness
3990    Kurtosis=K2*K2*Pdf(I)+Kurtosis
4000  NEXT I
4010  Var=Var/Numpoints
4020  Stdev=SQRT(Var)
```

```
4030  Skewness=Skewness/Numpoints/Stdev^3
4040  Kurtosis=Kurtosis/Numpoints/Var^2
4050  K3=-2*Stdev^2
4060  K4=Stdev*SQRT(2*PI)
4070  Ln2=LOG(2)
4080  Sumerr=0
4090  Entropy=0
4100  ALLOCATE Gauss(K+2)
4110  FOR I=0 TO K+1
4120    Gauss(I)=Numpoints*(EXP((I-Mean)^2/K3)/K4)
4130    Sumerr=ABS(Pdf(I)-Gauss(I))+Sumerr
4140    ON ERROR GOTO Off_error
4150    Entropy=Pdf(I)/Numpoints*LOG(Pdf(I)/Numpoints)/Ln2+Entropy
4160 Off_error:OFF ERROR
4170  NEXT I
4180  Entropy=LOG(Stdev*SQRT(2*PI*EXP(1)))/Ln2+Entropy
4190  Sumerr=Sumerr/Numpoints
4200  Low=INT(Mean-3*Stdev)
4210  IF Low<0 THEN Low=0
4220  High=INT(Mean+3*Stdev)
4230  IF High>(K+1) THEN High=K+1
4240  D=0
4250  D2=0
4260  M=High-Low+1
4270  FOR I=Low TO High
4280    Dtemp=ABS(Pdf(I)-Gauss(I))/Gauss(I)
4290    D=Dtemp+D
4300    D2=Dtemp^2+D2
4310  NEXT I
4320  Avgerr=D/M
4330  Rmserr=SQRT(D2/M)
4340  Term1=(Sumerr+Avgerr+Rmserr)/3
4350  Term2=ABS(Entropy)
4360  Term3=(ABS(Kurtosis-3)+ABS(Skewness))/2
4370  Reciprocal=(Term1+Term2+Term3)/3
4380  Quality=1.0/Reciprocal
4390  Quality=PROUND(Quality,-4)
4400  PRINT "TURNER NOISE QUALITY IS";Quality
4410  KEY LABELS ON
4420  STOP
4430  END
4440  DEF FNErf(X)                      ! COMPLEMENTARY ERROR FUNCTION
4450    Prob=0
4460    FOR N=1 TO 33 STEP 2
4470      Exp_arg=-(N*PI/(14*SQRT(2)))^2
4480      Sin_arg=(N*PI*X/14)
4490      Prob=Prob+((EXP(Exp_arg)*SIN(Sin_arg)/N))
4500    NEXT N
4510    Prob=(.5-2*Prob/PI)
4520    RETURN Prob
4530  FNEND
```

A.5 IF Noise Quality

Since this program depends on the statistical grouping of voltage samples along with spectrum analyzer measurements, the results of this program are dependent upon the number of voltage samples used to calculate the time-domain penalty and the spectrum analyzer settings, along with the SMOOTH parameter

(line 430), used to calculate the frequency-domain penalty. The total number of voltage samples used depends on the sampling rate, bandwidth of the filter or signal being measured, and the number of memory blocks downloaded to the computer. Note that the total number of samples used is not necessarily equal to the total number of samples acquired from the HP 54111D oscilloscope; correlated samples are ignored. Spectrum analyzer settings, particularly the video bandwidth, affect the calculation of the frequency-domain penalty. Finally, voltage samples are acquired from the HP 54111D with the RESOLUTION filter set to OFF for quicker data-acquisition times. Additional comments are imbedded in the code. Please read Chapter VI before using this program.

```
10    ! HP BASIC 5.14
20    !
30    ! PROGRAM IFNQ.BAS
40    !
50    ! THIS PROGRAM COMPUTES THE NOISE QUALITY OF A SIGNAL. ALTHOUGH
60    ! THE PROGRAM IS CALLED "IF NOISE QUALITY," IT CAN BE USED TO
70    ! MEASURE THE NOISE QUALITY OF ANY SIGNAL AT BASEBAND OR IF.
80    ! IF NOISE QUALITY IS CALCULATED BY ASSESSING A TIME-DOMAIN
90    ! PENALTY FOR NON-NORMALITY IN THE TIME-DOMAIN AND A FREQUENCY-
100   ! DOMAIN PENALTY FOR NON-FLATNESS OF THE POWER SPECTRUM IN THE
110   ! FREQUENCY-DOMAIN. MUCH OF THE CODE IN THIS PROGRAM, SUCH AS
120   ! DATA-ACQUISITION AND DATA-PROCESSING ALGORITHMS, ARE THE SAME
130   ! AS THOSE FOUND IN TURNER NOISE QUALITY. REFER TO THAT PROGRAM
140   ! LISTING FOR MORE INFORMATION.
150   CLEAR SCREEN
160   KEY LABELS ON
170   OPTION BASE 0
180   BEEP
190   OUTPUT 718;"CR;CV;CONTS;"          ! COUPLE VIDEO AND RESOLUTION
200   LOCAL 718                          ! BANDWIDTHS; CONTINUOUS SWEEP;
210                                      ! PUT ANALYZER IN LOCAL; COMPARATIVE
220                                      ! MEASUREMENTS SHOULD BE MADE WITH
230                                      ! SAME VIDEO BANDWIDTH
240                                      !
250                                      !
260   PRINT "SET OSCILLOSCOPE SETTINGS FOR SLOWEST SAMPLING RATE POSSIBLE."
270   PRINT ""
280   PRINT ""
290   PRINT "SET SPECTRUM ANALYZER SPAN WIDTH FOR 3 dB BANDWIDTH OF FILTER."
300   PRINT ""
310   PRINT ""
320   PRINT "SET OTHER SPECTRUM ANALYZER SETTINGS AS NECESSARY."
330   PRINT ""
340   PRINT ""
350   INPUT "WHAT IS THE BANDWIDTH OF YOUR FILTER IN HZ?",Bw
360   CLEAR SCREEN
370   Dumps=2
380   Numpoints=Dumps*8192
390   ALLOCATE Anoise(Numpoints-1)
400   ALLOCATE Asorted(Numpoints-1)
410   PRINT "SWEEPING TRACES AND ACQUIRING DATA."
420   PRINT ""
430   OUTPUT 718;"VBO -1;S2;TS;SMOOTH TRA 35;"
```

```
440   OUTPUT 707;"ACQUIRE TYPE NORMAL"
450   OUTPUT 707;"ACQUIRE RESOLUTION OFF"      ! DIGITIZE TIME-DOMAIN DATA
460   OUTPUT 707;"WAVEFORM SOURCE MEMORY 2"
470   OUTPUT 707;"WAVEFORM FORMAT ASCII"
480   M=0
490   N=0
500   OUTPUT 707;"BLANK CHANNEL 1"
510   OUTPUT 707;"BLANK CHANNEL 2"
520   REPEAT
530     IF Dumps>1 THEN PRINT TAB(10);"ACQUIRING TRACE ";M+1
540     Counter=0
550     OUTPUT 707;"DIGITIZE CHANNEL 2"
560     OUTPUT 707;"DATA?"
570     REPEAT
580       ENTER 707;Anoise(N)                      ! READ SCOPE DATA
590       Counter=Counter+1
600       N=N+1
610     UNTIL Counter=8191
620     M=M+1
630   UNTIL M=Dumps
640   OUTPUT 707;"YREFERENCE?"
650   ENTER 707;Yref
660   OUTPUT 707;"YINCREMENT?"
670   ENTER 707;Yinc
680   OUTPUT 707;"YORIGIN?"
690   ENTER 707;Yorg
700   OUTPUT 707;"XINCREMENT?"
710   ENTER 707;Xinc
720   Tstep=INT(1/(Bw*Xinc))+1
730   OUTPUT 707;"VIEW CHANNEL 1"
740   OUTPUT 707;"VIEW CHANNEL 2"
750   OUTPUT 707;"LOCAL"
760   Counter=0
770   FOR N=0 TO Numpoints-1 STEP Tstep
780     Anoise(N)=Anoise(N)
790     Asorted(N)=Anoise(N)
800     Counter=Counter+1
810   NEXT N
820   Numpoints=Counter
830   CLEAR SCREEN
840   PRINT "PERFORMING DIGITAL TO ANALOG CONVERSION."
850   FOR N=0 TO Numpoints-1
860     Anoise(N)=((Anoise(N)-Yref)*Yinc)+Yorg
870     Asorted(N)=Anoise(N)
880   NEXT N
890   ALLOCATE Noise(Numpoints-1)
900   ALLOCATE Sorted(Numpoints-1)
910   FOR N=0 TO Numpoints-1
920     Noise(N)=Anoise(N)
930     Sorted(N)=Asorted(N)
940   NEXT N
950   DEALLOCATE Anoise(*)
960   DEALLOCATE Asorted(*)
970   CLEAR SCREEN
980   PRINT "ESTIMATING PARAMETERS."
990   Xbar=0
1000  FOR N=0 TO Numpoints-1                       ! SAMPLE MEAN
1010    Xbar=Noise(N)+Xbar
1020  NEXT N
1030  Xbar=Xbar/(Numpoints)
1040  S2=0
1050  FOR N=0 TO Numpoints-1
1060    S2=S2+(Noise(N)-Xbar)^2
```

```
1070  NEXT N
1080  S2=S2/(Numpoints-1)                              ! SAMPLE VARIANCE
1090  S=SQRT(S2)
1100  FOR N=0 TO Numpoints-1                   ! NORMARLIZE TO ZERO-MEAN
1110    Noise(N)=(Noise(N)-Xbar)/S             ! AND UNITY VARIANCE
1120    Sorted(N)=(Sorted(N)-Xbar)/S
1130  NEXT N
1140  Left=MIN(Noise(*))
1150  Right=MAX(Noise(*))
1160  MAT SORT Sorted(*)                 ! DETERMINE START AND END POINTS
1170  Amark=2                            ! FOR K EQUAL WIDTH BINS
1180  A=Sorted(Amark)                    ! AND SORT VOLTAGES INTO BINS
1190  Bmark=Numpoints-2
1200  B=Sorted(Bmark)
1210  IF (A=Left) THEN
1220    Counter=Amark                    ! FORCE LEFT-END AND RIGHT-END
1230    REPEAT                           ! BINS TO CONTAIN AT LEAST TWO
1240      Counter=Counter+1              ! SAMPLES AND BE OF NON-ZERO WIDTH
1250      A=Sorted(Counter)
1260    UNTIL A>Sorted(Counter-1)
1270  END IF
1280  IF (B=Right) THEN
1290    Counter=Bmark
1300    REPEAT
1310      Counter=Counter-1
1320      B=Sorted(Counter)
1330    UNTIL B<Sorted(Counter+1)
1340  END IF
1350  Delta=.2
1360  K=(B-A)/Delta
1370  K=INT(K)
1380  Delta=(B-A)/K
1390  ALLOCATE Pdf(K+1)
1400  CLEAR SCREEN
1410  PRINT "SORTING VOLTAGES INTO BINS."
1420  FOR M=0 TO Numpoints-1
1430    SELECT Noise(M)
1440    CASE <A                                      ! GROUP SAMPLES INTO BINS
1450      Pdf(0)=Pdf(0)+1
1460    CASE >=B
1470      Pdf(K+1)=Pdf(K+1)+1
1480    CASE ELSE
1490      J=INT((Noise(M)-A)/Delta)+1
1500      Pdf(J)=Pdf(J)+1
1510    END SELECT
1520  NEXT M
1530  CLEAR SCREEN
1540  Dxbar=PROUND(Xbar,-4)                          ! PRINT INFO
1550  Ds=PROUND(S,-4)
1560  Dl=PROUND(Left,-4)
1570  Dr=PROUND(Right,-4)
1580  Mf=MIN(Pdf(*))
1590  Xf=MAX(Pdf(*))
1600  CLEAR SCREEN
1610  PRINT "SAMPLE MEAN =";Dxbar,TAB(40),"SAMPLE STANDARD DEVIATION =";Ds
1620  PRINT ""
1630  PRINT "MINIMUM VOLTAGE = ";Dl,TAB(40),"MAXIMUM VOLTAGE =";Dr
1640  PRINT ""
1650  PRINT "MINIMUM FREQUENCY =";Mf,TAB(40),"MAXIMUM FREQUENCY =";Xf
1660  PRINT ""
1670  PRINT "NUMBER OF POINTS = ";SUM(Pdf),TAB(40),"NUMBER OF INTERVALS = ";K+2
1680  PRINT ""
1690  PRINT "BIN WIDTH = ";PROUND(Delta*S,-4)
```

```
1700  PRINT ""
1710  PRINT "NOISE CORRELATION DURATION GTE TO ";1/(Bw)
1720  PRINT ""
1730  PRINT "TIME BETWEEN DECORRELATED SAMPLES IS APPROX ";Tstep*Xinc
1740  PRINT ""
1750  PRINT ""
1760  PRINT ""
1770  PRINT TAB(25),"Press F2 to CONTINUE"
1780  PAUSE
1790  CLEAR SCREEN
1800  PRINT "CALCULATING TIME-DOMAIN PENALTY."
1810  WAIT 2
1820  Low=A
1830  Nrmse=0
1840  ALLOCATE Ideal(K)
1850  FOR I=1 TO K                          ! CALCULATE AVERAGE NORMALIZED
1860    Hi=A+I*Delta                        ! RMS ERROR OF CALCULATED PDF
1870    X_ci=(Hi+Low)/2                     ! AND COMPARE TO NORMAL PDF
1880    Exp_arg=-.5*(X_ci)^2
1890    Fx_ci=EXP(Exp_arg)/SQRT(2*PI)
1900    Ideal(I)=Fx_ci*Delta*Numpoints
1910    Obs=Pdf(I)/(Delta*Numpoints)
1920    Nrmse=Nrmse+SQRT((Obs-Fx_ci)^2/Fx_ci^2)
1930    Low=Hi
1940  NEXT I
1950  Nrmse=Nrmse/K                          ! TIME-DOMAIN PENALTY
1960  CLEAR SCREEN
1970  Left=MIN(Noise(*))
1980  Right=MAX(Noise(*))
1990  GINIT                                  ! PLOT THEORETICAL
2000  GCLEAR                                 ! PDF (HISTOGRAM)
2010  GRAPHICS ON
2020  KEY LABELS OFF
2030  MOVE 27,90
2040  LABEL "     IDEAL HISTOGRAM"
2050  MOVE 0,70
2060  Label$="Frequency"
2070  FOR I=1 TO 9
2080    LABEL Label$[I,I]
2090  NEXT I
2100  MOVE 55,10
2110  LABEL "Volts"
2120  VIEWPORT 10,120,15,90
2130  FRAME
2140  WINDOW Left,Right,0,MAX(Ideal(*))+5
2150  FOR I=0 TO K+1
2160    SELECT I
2170    CASE =0
2180      MOVE Left,0
2190      RECTANGLE ABS(Left-A),Pdf(0)
2200    CASE =K+1
2210      MOVE B,0
2220      RECTANGLE ABS(Right-B),Pdf(K+1)
2230    CASE ELSE
2240      MOVE A+(I-1)*Delta,0
2250      RECTANGLE Delta,Ideal(I)
2260    END SELECT
2270  NEXT I
2280  PRINT "PRESS F2 TO CONTINUE."
2290  PAUSE
2300  CLEAR SCREEN
2310  KEY LABELS ON
2320  CLEAR SCREEN
```

```
2330 GINIT                                   ! PLOT HISTOGRAM
2340 GCLEAR                                  ! (PDF ESTIMATE)
2350 GRAPHICS ON                             ! OF DATA
2360 KEY LABELS OFF
2370 MOVE 27,90
2380 LABEL "HISTOGRAM OF NOISE VOLTAGES"
2390 MOVE 0,70
2400 Label$="Frequency"
2410 FOR I=1 TO 9
2420   LABEL Label$[I,I]
2430 NEXT I
2440 MOVE 55,10
2450 LABEL "Volts"
2460 VIEWPORT 10,120,15,90
2470 FRAME
2480 WINDOW Left,Right,0,MAX(Pdf(*))+5
2490 FOR I=0 TO K+1
2500   SELECT I
2510   CASE =0
2520     MOVE Left,0
2530     RECTANGLE ABS(Left-A),Pdf(0)
2540   CASE =K+1
2550     MOVE B,0
2560     RECTANGLE ABS(Right-B),Pdf(K+1)
2570   CASE ELSE
2580     MOVE A+(I-1)*Delta,0
2590     RECTANGLE Delta,Pdf(I)
2600   END SELECT
2610 NEXT I
2620 PRINT "PRESS F2 TO CONTINUE."
2630 PAUSE
2640 CLEAR SCREEN
2650 KEY LABELS ON
2660 ALLOCATE Expected(K+1)
2670 FOR I=0 TO K+1
2680   Prob=0                                ! CALCULATE EXPECTED
2690   SELECT I                              ! FREQUENCIES FOR
2700   CASE =0                               ! CHI_SQUARE TEST.
2710                                         ! TEST WILL BE PERFORMED
2720     X=A                                 ! WITH K+2 BINS.  HENCE,
2730     Prob=FNErf(X)                       ! K-1 DEGREES OF FREEDOM.
2740     Expected(0)=Numpoints*(1-Prob)
2750   CASE =K+1
2760     X=B
2770     Prob=FNErf(X)
2780     Expected(K+1)=Numpoints*Prob
2790   CASE ELSE
2800     Low_lim=A+((I-1)*Delta)
2810     Up_lim=A+(I*Delta)
2820     Prob=FNErf(Low_lim)-FNErf(Up_lim)
2830     Expected(I)=Numpoints*Prob
2840   END SELECT
2850 NEXT I
2860 DIM Chiray(29)
2870 DATA 3.84,5.99,7.81,9.49,11.07,12.59,14.07,15.51,16.92,18.31
2880 FOR N=0 TO 9
2890   READ Chiray(N)
2900 NEXT N
2910 DATA 19.68,21.03,22.36,23.68,25,26.30,27.59,28.87,30.14,31.41
2920 FOR N=10 TO 19
2930   READ Chiray(N)
2940 NEXT N
2950 DATA 32.67,33.92,35.17,36.42,37.65,38.88,40.11,41.34,42.56,43.77
```

```
2960 FOR N=20 TO 29
2970   READ Chiray(N)
2980 NEXT N
2990 Chi_square=0                      ! PERFORM CHI-SQUARE TEST AT
3000                                   ! a = .05 LEVEL OF SIGNIFCANCE
3010                                   ! INCLUDE YATE'S CORRECTION
3020 FOR I=0 TO K+1
3030   Term=ABS(Pdf(I)-Expected(I))
3040   Chi_square=Chi_square+((Term-.5)^2)/Expected(I)
3050 NEXT I
3060 Chi_square=PROUND(Chi_square,-2)
3070 IF K-1<=29 THEN
3080   X2=Chiray(K-2)
3090 ELSE
3100   X2=.5*(SQRT(2*(K-1))+1.645)^2
3110 END IF
3120 PRINT "IDEAL CHI-SQUARE = ";PROUND(X2,-2)
3130 PRINT ""
3140 PRINT "CALCULATED CHI-SQUARE = ";Chi_square
3150 PRINT ""
3160 IF (Chi_square<=X2) THEN
3170   PRINT "CHI-SQUARE DOES NOT REJECT THE HYPOTHESIS OF NORMALITY."
3180 ELSE
3190   PRINT "CHI-SQUARE REJECTS THE HYPOTHESIS OF NORMALITY."
3200 END IF
3210 PRINT ""
3220 PRINT ""
3230 PRINT "PRESS F2 TO CONTINUE."
3240 PAUSE
3250 CLEAR SCREEN
3260 PRINT "CALCULATING FREQUENCY-DOMAIN PENALTY."
3270 ALLOCATE Psd(1000)                    ! ACQUIRE DATA FROM ANALYZER
3280 OUTPUT 718;"TA;"
3290 ENTER 718;Psd(*)
3300 Tot_power=0
3310 FOR I=0 TO 1000
3320   Tot_power=Tot_power+10^(Psd(I)/10)   ! CALCULATE POWER RATIO
3330 NEXT I
3340 Max_dbm=MAX(Psd(*))
3350 Flat_power=1001*10^(Max_dbm/10)
3360 Pwr_ratio=Tot_power/Flat_power
3370 CLEAR SCREEN
3380 PRINT "TIME-DOMAIN PENALTY IS ";PROUND(Nrmse,-4)
3390 PRINT ""
3400 PRINT "FREQUENCY-DOMAIN PENALTY IS ";PROUND(Pwr_ratio,-4)
3410 PRINT ""
3420 Ifnq=PROUND((1-Nrmse)*Pwr_ratio,-3)
3430 PRINT "IF NOISE QUALITY IS ";Ifnq*100;"%"        ! IF NOISE QUALITY
3440 LOCAL 718                     ! PUT SPECTRUM ANALYZER IN LOCAL
3450 STOP
3460 END
3470 DEF FNErf(X)                  ! COMPLEMENTARY ERROR FUNCTION
3480   Prob=0
3490   FOR N=1 TO 33 STEP 2
3500     Exp_arg=-(N*PI/(14*SQRT(2)))^2
3510     Sin_arg=(N*PI*X/14)
3520     Prob=Prob+((EXP(Exp_arg)*SIN(Sin_arg)/N))
3530   NEXT N
3540   Prob=(.5-2*Prob/PI)
3550   RETURN Prob
3560 FNEND
```

A.6 RF Noise Quality

Since this program depends on spectrum analyzer measurements, the results of this program are dependent upon the spectrum analyzer settings, along with the SMOOTH parameter (line 430), used to measure the baseband normality and the frequency-domain impulsivity. Spectrum analyzer settings, particularly the video bandwidth, affect the calculation of the time- and frequency-domain penalties. Additional comments are imbedded in the code. Please read Chapter VI before using this program.

```
10    !  HP BASIC 5.14
20    !
30    !  PROGRAM RFBNQ.BAS
40    !
50    !  THIS PROGRAM CALCULATES THE BARRAGE NOISE QUALITY
60    !  OF AN RF FM-BY-NOISE BARRAGE BASED ON DATA ACQUIRED
70    !  FROM AN HP 8566B SPECTRUM ANALYZER (ADDRESS 718).
80    !
90    OPTION BASE 0
100   OUTPUT 718; "CR; CV; "        ! PUT RESOLUTION AND VIDEO BANDWIDTHS ON AUTO
110   CLEAR SCREEN
120   LOCAL 718                                  ! PUT ANALYZER IN LOCAL MODE
130   INPUT "WHAT IS THE PEAK FREQUENCY DEVIATION IN HZ?  0 =  UNKNOWN",Pfd
140   CLEAR SCREEN
150   IF (Pfd = 0) THEN
160     BEEP
170     INPUT "WHAT IS THE BARRAGE WIDTH IN HZ?",Bw
180     CLEAR SCREEN
190   END IF
200   PRINT "CENTER UNMODULATED CARRIER ON SPECTRUM ANANLYZER."
210   PRINT "LOCK SIGNAL GENERATOR IF POSSIBLE."
220   PRINT ""
230   PRINT "AFTER UNMODULATED CARRIER IS CENTERED, BEGIN JAMMING."
240   PRINT ""
250   PRINT "PRESS F2 TO CONTINUE"
260   PAUSE
270   IF (Pfd = 0) THEN
280     OUTPUT 718; "SP "&VAL$(Bw)&"; "                        ! SET SPAN WIDTH
290   ELSE                                                     ! FOR BARRAGE
300     OUTPUT 718; "SP "&VAL$(2*Pfd)&"; "
310   END IF
320   CLEAR SCREEN
330   BEEP
340   LOCAL 718
350   PRINT "SET dB/DIV AND REFERENCE LEVEL FOR MAXIMUM RESOLUTION"
360   PRINT "AND"
370   PRINT "RESET VIDEO BANDWIDTH FOR A SMOOTH TRACE."
380   PRINT ""
390   PRINT ""
400   PRINT ""
410   PRINT "PRESS F2 TO CONTINUE"
420   PAUSE
430   !
440   ! BARRAGE COMPARISONS BASED ON DIFFERENT BASEBAND NOISE BANDWIDTHS
450   ! SHOULD BE MADE WITH THE SAME VIDEO BANDWIDTH SETTING. NOISE
460   ! AVERAGING IS INVERSELY PROPORTIONAL TO VIDEO BANDWIDTH, WHILE
470   ! SPEED IS PROPORTIONAL TO VIDEO BANDWIDTH. WIDE VIDEO BANDWIDTHS
480   ! GIVE QUICK MEASUREMENTS WITH LITTLE AVERAGING OF THE NOISE ON
```

```
490   ! ON THE TRACE. NARROW VIDEO BANDWIDTHS PRODUCE AVERAGED TRACES
500   ! BUT CAUSE THE SWEEP SPEED OF THE TRACE TO SLOW DOWN.
510   ! TRACES BUT CAUSE THE SWEEP SPEED OF THE TRACE TO SLOW DOWN.
520   ! THE FIRST PART OF THE PROGRAM MEASURES THE NOISE QUALITY IN TERMS
530   ! OF WOODWARD'S THEOREM AND BASEBAND NOISE THAT IS ASSUMED TO BE
540   ! GAUSSIAN. IF THE BASEBAND NOISE IS GAUSSIAN, THEN, ACCORDING TO
550   ! WOODWARD'S THEOREM THE RESULTING RF BARRAGE SHOULD BE BELL-SHAPED.
560   ! WOODWARD'S THEOREM IS APPLICABLE TO WIDEBAND FM WHICH IS ASSUMED AT
570   ! THE OUTSET. NARROW VIDEO BANDWIDTHS ARE BEST FOR INVESTIGATING THE
580   ! SHAPE OF A BARRAGE. HENCE, CHOOSE THE NARROWEST VIDEO BANDWIDTH
590   ! TOLERABLE. ALSO NOTE THAT THE IDEAL JAMMING POWER IS BASELINED TO
600   ! THE MAXIMUM MAGNITUDE OF THE TRACE WHEN THE FULL BARRAGE IS DIS-
610   ! PLAYED TRACE MAGNITUDES IS DEPENDENT UPON THE RESOLUTION BANDWIDTH
620   ! SELECTED WHEN THE TRACE IS MEASURED. HENCE, THE RESOLUTION BANDWIDTH
630   ! IS QUERIED AND USED TO MAKE THE WHITE-NOISE MEASUREMENT. OTHERWISE
640   ! THE FREQUENCY DOMAIN PENALTY WOULD BE ARTIFICIALLY HIGH.
650   !
660   CLEAR SCREEN                                ! ANALYZER UNITS KSA =  dBm
670   OUTPUT 718; "KSA;  SAVES 6; "               ! SAVE THE INSTRUMENT SET-UP
680   DIM S(-500:500)                             ! DIMENSION AN ARRAY
690   OUTPUT 718; "KSD; S2,TS; O3; "              ! OUTPUT VOLTS, SINGLE
700   OUTPUT 718; "SMOOTH TRA 17; "               ! SWEEP,SWEEP,ASCII
710   OUTPUT 718; "TA; "                          ! SMOOTH TRACE,
720   ENTER 718; S(*)                             ! DOWNLOAD TRACE
730   S2 = 0
740   Limit = 500                                 ! DETERMINES NUMBER OF POINTS
750   Temp = 0                                    ! TO USE FOR CURVE-FITTING
760   Counter = 0
770   FOR N = -Limit TO Limit                     ! CURVE-FITTING TO A
780     IF S(N) >= S(0) THEN GOTO 830             ! NORMAL CURVE;  ASSUME
790     Ln_term = LOG(S(N)/S(0))                  ! "ZERO-MEAN" AND ESTIMATE
800     Temp = -.5*N^2/Ln_term                    ! "VARIANCE"
810     S2 = S2+Temp
820     Counter = Counter+1
830   NEXT N
840   S2 = S2/(Counter-1)
850   Nrmse = 0
860   FOR N = -Limit TO Limit                     ! CALCULATE NORMALIZED
870     Theta = EXP(-.5*(N)^2/S2)*S(0)            ! RMS ERROR
880     Temp = SQRT(((S(N)-Theta)^2)/Theta^2)
890     Nrmse = Nrmse +Temp
900   NEXT N
910   Nrmse = Nrmse/(2*Limit+1)             ! NORMLIZED RMS ERROR
920   OUTPUT 718; "KSA; MKPK; "
930   OUTPUT 718; "MA; "
940   ENTER 718; Max_dbm                          ! READ IN VALUE TO BASELINE
950   OUTPUT 718; "MKOFF; RB?"                    ! IDEAL JAMMING POWER AND
960   ENTER 718; Res_bw                           ! SAVE RESOLUTION BANDWIDTH
970   IF Pfd = 0 THEN
980     OUTPUT 718; "M2; "                        ! SET UP ANALYZER
990     OUTPUT 718; "MKD,MKTYPE AMP, MKA -3; "    ! FOR THE 3dB BW
1000    OUTPUT 718; "M3?"                         ! OF THE BARRAGE
1010    ENTER 718; Delta_f                        ! FOR UNKNOWN PEAK
1020    OUTPUT 718; "MKOFF; "                     ! FREQ DEVIATION
1030    OUTPUT 718; "CF?"
1040    ENTER 718; Cf
1050    F_a = Cf-ABS(Delta_f)
1060    F_b = Cf+ABS(Delta_f)
1070    F_a$ = VAL$(F_a)
1080    F_b$ = VAL$(F_b)
1090    OUTPUT 718; "FA "&F_a$&"; "
1100    OUTPUT 718; "FB "&F_b$&"; "
1110  ELSE
```

```
1120    Span = .8*Pfd                         ! ESTIMATE 3 dB BW BASED
1130    Span$ = VAL$(Span)                    ! ON PEAK FREQ DEVIATION
1140    OUTPUT 718; "SP "&Span$&"; "
1150  END IF
1160  !
1170  ! NOTE: THE 3 dB BANDWIDTH ESTIMATE ASSUMES THE STATISTICS OF THE
1180  ! BASEBAND MODULATING NOISE ARE APPROXIMATELY NORMAL WITH ZERO
1190  ! MEAN AND VARIANCE, SIGMA^2 OR JUST S^2.  THE 3 dB BW OF AN
1200  ! FM-BY-NOISE BARRAGE MODULATED BY GAUSSIAN NOISE HAS BEEN SHOWN
1210  ! TO BE APPROXIMATELY 2.36 * (INSTANTANEOUS RMS FREQUENCY DEVIATION).
1220  ! IN THE CASE OF A CARRIER BEING FREQUENCY MODULATED BY N(0,S^2)
1230  ! NOISE, THE PEAK FREQUENCY DEVIATION IS APPROXIMATELY EQUAL TO
1240  ! (PEAK DEVIATION CONSTANT, IN HZ/VOLT)*(3*SIGMA).  IN OTHER WORDS,
1250  ! THE PEAK DEVIATION IS EQUAL TO THE PEAK DEVIATION CONSTANT TIMES
1260  ! THE MAXIMUM VOLTAGE, WHICH IS APPROXIMATELY 3*SIGMA.  HENCE, THE
1270  ! INSTANTANEOUS RMS FREQUENCY DEVIATION IS APPROXIMATELY EQUAL TO
1280  ! (2.36/3)*(PEAK DEVIATION) FOR NORMAL NOISE.
1290  !
1300  OUTPUT 718; "KSA; RB "&VAL$(Res_bw)&"; S2; TS; SMOOTH TRA 17; "
1310  OUTPUT 718; "TA; "
1320  ENTER 718; S(*)
1330  Tot_power = 0                                    ! CALCULATE RATIO
1340  FOR I = -500 TO 500                              ! OF MEASURED POWER
1350    Tot_power = Tot_power+10^(S(I)/10)             ! TO IDEAL FLAT POWER
1360  NEXT I                                           ! IN THE BARRAGE'S
1370                                                   ! 3 dB BW
1380  Flat_power = 1001*10^(Max_dbm/10)
1390  Pwr_ratio = Tot_power/Flat_power
1400  PRINT "NORMALIZED RMS ERROR IS",PROUND(Nrmse ,-4)
1410  PRINT "POWER RATIO IS",PROUND(Pwr_ratio,-4)              ! CALCULATE
1420  Bnq = PROUND((1-Nrmse )*Pwr_ratio,-3)                    ! RF NOISE
1430  PRINT "RF NOISE QUALITY IS "; Bnq*100; "%"               ! QUALITY
1440  OUTPUT 718; "RCLS 6; "
1450  LOCAL 718
1460  END
```

Appendix B Equipment and Software

Table B-1 lists the equipment and software used to conduct the experimental portion of this investigation. Table B-2 describes the specifications of the HP 54111D digitizing oscilloscope in terms of sweep speed, sampling rate, memory depth, vertical resolution, and bandwidth. Vertical resolution was selected by setting the vertical RESOLUTION filters to OFF, 6 bits, 7 bits, or 8 bits. With the vertical RESOLUTION filters set to OFF, the bandwidth of the HP 54111D was 1/2 the sampling rate.

As mentioned in Chapter IV, other equipment was tested but not used; this equipment is not included in the table. Additionally, an operations manual accompanied each piece of equipment and software listed in the table. However, the manuals are not included in the Bibliography. Their inclusion would have made the Bibliography unwieldy.

Table B-1 Table of Equipment and Software

SIMULATED JAMMER		
ITEM	*COMPANY*	*MODEL*
Noise Generator	Hewlett Packard Co.	HP 3722A
Signal Generator	Hewlett Packard Co.	HP 8640B
SIMULATED VICTIM RECEIVER		
ITEM	*COMPANY*	*MODEL*
Mixer	Anzac	MD 141
Signal Generator	Hewlett Packard Co.	HP 8640B
Dual Hi/Lo Filter	Wavetek Rockland	Model 852
MEASUREMENT EQUIPMENT		
ITEM	*COMPANY*	*MODEL*
Oscilloscope	Hewlett Packard Co.	HP 54111D
Spectrum Analyzer	Hewlett Packard Co.	HP 8566B
Computer	IBM	286 PC
Coprocessor	Hewlett Packard Co.	82324A
Software	Hewlett Packard Co.	HP Basic 5.14

Table B-2 HP 54111D Digitizing Oscilloscope Specifications

Sweep Speed	Sample Rate	Memory Depth	Bandwidth		
			Vertical Resolution		
			6 bits	7 bits	8 bits
500 ps/div - 99.9 ns/div	1 gigasamples/s	8.19 µs	250 MHz	100 MHz	25 MHz
100 ns/div - 199 ns/div	500 megasamples/s	16.3 µs	125 MHz	50 MHz	12.5 MHz
200 ns/div - 499 ns/div	250 megasamples/s	32.7 µs	62.5 MHz	25 MHz	6.25 MHz
500 ns/div - 999 ns/div	100 megasamples/s	81.9 µs	25 MHz	10 MHz	2.5 MHz
1 µs/div - 1.99 µs/div	50 megasamples/s	163 µs	12.5 MHz	5 MHz	1.25 MHz
2 µs/div - 4.99 µs/div	25 megasamples/s	327 µs	6.25 MHz	2.5 MHz	625 kHz
5 µs/div - 9.99 µs/div	10 megasamples/s	819 µs	2.5 MHz	1 MHz	250 kHz
10 µs/div - 19.9 µs/div	5 megasamples/s	1.63 ms	1.25 MHz	500 kHz	125 kHz
20 µs/div - 49.9 µs/div	2.5 megasamples/s	3.27 ms	625 kHz	250 kHz	62.5 kHz
50 µs/div - 99.9 µs/div	1 megasamples/s	8.19 ms	250 kHz	100 kHz	25 kHz
100 µs/div - 199 µs/div	500 kilosamples/s	16.3 ms	125 kHz	50 kHz	12.5 kHz
200 µs/div - 499 µs/div	250 kilosamples/s	32.7 ms	62.5 kHz	25 kHz	6.25 kHz
500 µs/div - 999 µs/div	100 kilosamples/s	81.9 ms	25 kHz	10 kHz	2.5 kHz
1 ms/div - 1.99 ms/div	50 kilosamples/s	163 ms	12.5Hz	5 kHz	1.25 kHz
2 ms/div - 4.99 ms/div	25 kilosamples/s	327 ms	6.25 kHz	2.5 kHz	625 Hz
5 ms/div - 9.99 ms/div	10 kilosamples/s	819 ms	2.5 kHz	1 kHz	250 Hz
10 ms/div - 19.9 ms/div	5 kilosamples/s	1.63 s	1.25 kHz	500 Hz	125 Hz
20 ms/div - 49.9 ms/div	2.5 kilosamples/s	3.27 s	625 Hz	250 Hz	62.5 Hz
50 ms/div - 99.9 ms/div	1 kilosamples/s	8.19 s	250 Hz	100 Hz	25 Hz
100 ms/div - 199 ms/div	500 samples/s	16.3 s	125 Hz	50 Hz	12.5 Hz
200 ms/div - 499 ms/div	250 samples/s	32.7 s	62.5 Hz	25 Hz	6.25 Hz
500 ms/div - 999 ms/div	100 samples/s	81.9 s	25 Hz	10 Hz	2.5 Hz
1 s/div	50 samples/s	163 s	6.25 Hz	5 Hz	1.25 Hz

Appendix C Results from Noise Quality Measurements

This appendix presents results obtained from preliminary tests of the noise quality measures developed in this investigation. The tables contain raw data; time constraints prevented a rigorous statistical analysis of the data. Trends evident in the data are commented upon in Chapter VI.

C.1 Turner Noise Quality

Tables C-1 and C-2 present the results obtained from preliminary tests of the Turner noise quality (TNQ) measure. χ^2 is the theoretical chi-square value, and X^2 is the measured chi-square value. The sampling rate of the HP 54111D was set at 500 kilosamples/sec with RESOLUTION filters OFF; Channel Two voltage sensitivity was set at 80 mV/div. The noise generator was set for a 1 V_{RMS} Gaussian output with INFINITE SEQUENCE LENGTH. See Chapter IV for more on equipment settings.

Table C-1 Turner Noise Quality for $B_V = 25$ kHz and $\Delta f_p = 150$ kHz

B_N (kHz)	*SVR*	*Samples (N)*	*Bins (K + 2)*	χ^2	X^2	*TNQ*
0.5	0.12	781	49	63.13	9564	0.7348
1.5	0.36	781	47	60.78	11250	0.4253
5	1.2	781	30	40.11	545	1.6014
15	3.6	781	30	40.11	107	3.6129
50	12	781	26	35.17	30.14	6.5410

Table C-2 Turner Noise Quality with $B_V = 50$ kHz and $\Delta f_p = 150$ kHz

B_N (kHz)	*SVR*	*Samples (N)*	*Bins (K + 2)*	χ^2	X^2	*TNQ*
0.5	0.03	1490	23	31.41	1025	2.4377
1.5	0.09	1490	32	42.56	1374	1.1145
5	0.30	1490	34	41.34	1116	1.1983
15	0.90	1490	31	40.11	134	4.5460
50	3.0	1490	27	36.42	34.52	6.6348

C.2 IF Noise Quality

Tables C-3 and C-4 present the results from preliminary tests of the IF noise quality (IFNQ) measure. χ^2 is the theoretical chi-square value, and X^2 is the measured chi-square value. The time-and frequency-domain penalties are p_t and p_f, respectively. The sampling rate of the HP 54111D was set at 500 kilosamples/sec with RESOLUTION filters OFF; Channel Two voltage sensitivity was set at 80 mV/div. All HP 8566B spectrum analyzer functions were COUPLED, i.e. on AUTO, with the exception of the video bandwidth which was set at 100 Hz for Table C-3 measurements and at 300 Hz for Table C-4 measurements. The argument of the HP 8566B SMOOTH function (line 430 of the IF noise quality program) was set at 35. The noise generator was set for a 1 V_{RMS} Gaussian output with INFINITE SEQUENCE LENGTH. See Chapter IV for more on equipment settings.

Table C-3 IF Noise Quality with $B_V = 25$ kHz and $\Delta f_p = 150$ kHz

B_N (kHz)	*SVR*	*Samples*	*Bins*	χ^2	X^2	p_t	p_f	*IFNQ*
0.5	0.12	781	46	59.61	1935	20.429	0.6202	-1205%
1.5	0.36	781	40	52.50	1246	4.5314	0.5329	-188.2%
5	1.2	781	34	45.31	475	1.0579	0.5801	-3.4%
15	3.6	781	31	41.34	112	0.4778	0.5169	27%
50	12	781	27	36.42	29.75	0.2353	0.7292	55.8%

Table C-4 IF Noise Quality with $B_V = 50$ kHz and $\Delta f_p = 150$ kHz

B_N (kHz)	*SVR*	*Samples*	*Bins*	χ^2	X^2	p_t	p_f	*IFNQ*
0.5	0.03	1490	24	59.61	2235	0.997	0.343	0.1%
1.5	0.09	1490	31	52.50	1583	0.8864	0.5705	6.5%
5	0.30	1490	33	45.31	1199	0.7123	0.5678	16.3%
15	0.90	1490	33	41.34	274	0.5222	0.6873	32.8%
50	3.0	1490	27	36.42	28.25	0.1272	0.5281	46.1%

C.3 RF Noise Quality

Tables C-5 and C-7 present the results from preliminary tests of the RF noise quality measure proposed in Chapter VI. The time- and frequency-domain penalties are labeled p_t and p_f, respectively. The RF noise quality of three barrages was measured three times; the measurements are labeled *Run 1* through *Run 3* in the tables. The peak frequency deviation in all cases was 150 kHz. All measurements were made with the following initial spectrum analyzer settings: span width = 300 kHz, resolution bandwidth = 300 kHz, video bandwidth = 10 Hz, and sweep = 48 s. The argument of the HP 8566B SMOOTH function was set at 17 (lines 100 and 1300 of the RF noise quality program). The carrier frequency was 250 MHz, and the noise generator was set for a 1 V_{RMS} Gaussian output with INFINITE SEQUENCE LENGTH. See Chapter IV for more on equipment settings.

Table C-5 RF Noise Quality with B_N = 500 Hz

Δf_p = 150 kHz	p_t	p_f	*RF Noise Quality*
Run 1	0.2964	0.5352	37.7%
Run 2	0.3996	0.3568	21.4%
Run 3	0.5838	0.5553	23.1%

Table C-6 RF Noise Quality with B_N = 5 kHz

Δf_p = 150 kHz	p_t	p_f	*RF Noise Quality*
Run 1	0.2690	0.7610	55.6%
Run 2	0.2371	0.7397	56.4%
Run 3	0.2517	0.7606	56.9%

Table C-7 RF Noise Quality with B_N = 50 kHz

Δf_p = 150 kHz	p_t	p_f	*RF Noise Quality*
Run 1	0.0381	0.7967	76.6%
Run 2	0.0136	0.7861	77.5%
Run 3	0.0259	0.7865	76.6%

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Captain Charles J. Daly was born on 7 October 1957 in Philadelphia, Pennsylvania. He graduated from West Catholic High School for Boys in 1975. Upon graduation from high school, he enlisted in the U.S. Air Force. After completing basic training in September 1975, he attended the Defense Language Institute and studied Chinese-Mandarin. After his language training, he served as a Chinese Cryptologic Linguist at the 6903rd Electronic Security Group (ESG), Osan Air Base, Republic of Korea from December 1976 to October 1982. He returned to the DLI in October 1982 and studied Korean. Subsequently, he returned to the 6903rd ESG and served as a Korean Cryptologic Linguist from January 1983 to September 1985. In September 1985, then Technical Sergeant Daly was accepted into the Air Force's Airmen's Education and Commissioning Program. He was sent to Texas A&M University, College Station, Texas, to major in Electrical Engineering. He graduated Magna Cum Laude in May 1988 and was commissioned a Second Lieutenant in the U.S. Air Force in September 1988. After his commissioning, he served as a Technical Engineer for the Technical Engineering Section of the 394th Test and Maintenance Squadron, Vandenberg Air Force Base, California. There he was responsible for the engineering issues associated with Strategic Air Command's Follow-on Test and Evaluation Program of Minuteman and Peacekeeper missiles. He came to the Air Force Institute of Technology in May 1991.

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13. ABSTRACT *(Maximum 200 words)* Among the jamming waveforms used in Electronic Warfare, FM-by-noise is the least understood, and no exhaustive analytical and experimental investigation of the subject exists. This investigation presents a thorough review, consolidation, and elucidation of the theory of FM-by-noise. To explain and predict the behavior of FM-by-noise at RF and IF, three useful ratios, namely the deviation-to-victim ratio, the noise-to-victim ratio, and the swwep-to-victim ratio, are developed. To complement the theory, results of FM-by-noise jamming experiments conducted using commercial test and measurement equipment are presented. Specifically, the time- and frequency-domain behavior of FM-by-noise at RF and IF is studied with the baseband noise bandwidth, peak frequency deviation, and victim receiver bandwidth as parameters. An important feature of the experimental portion of this investigation is the inclusion of reproductions of actual oscilloscope and spectrum analyzer displays illustrating the behavior of FM-by-noise. Finally, the concept of noise quality, as a measure of noise jammer effectiveness, is reexamined. Three noise quality factors are investigated. An existing noise quality measure is verified, and two alternative noise quality factors are proposed.

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